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Simplified Dual-standard Video Detector and A.G.C. System

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The switching operations in a dual-standard television receiver may be minimised if the video detector circuit can produce a signal of similar polarity and d.c. level on both standards. A simple double-diode detector circuit is described that enables this to be achieved and at the same time is free from blocking effects when sudden signal level changes are made.

INTRODUCTION

In a dual-standard receiver many switching operations are necessary, particularly in the vision i.f. amplifier. The video detector output is reversed with a double-pole switch to obtain a video waveform of similar polarity with both positive (405-line) and negative (625-line) modulation. To minimise switching operations in the video amplifier and a.g.c. system, resulting from the difference in d.c. level as illustrated in Fig. 1, a.c. coupling in the following circuits is often employed.

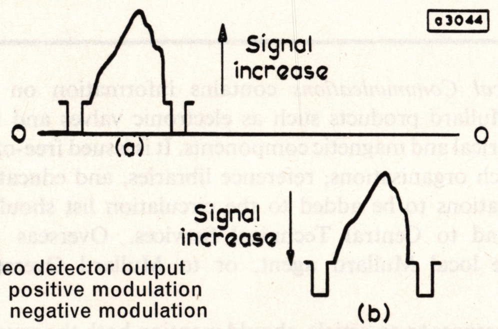

Fig. 1—Video detector output
(a) positive modulation
(b) negative modulation

When preservation of the d.c. level following the video detector is important, it may be re-inserted with a d.c. restoration diode. To prevent noise peaks on the 405-line system (Bands I and II) from driving the video amplifier into grid current and producing a spurious charge on the coupling capacitor, nominal d.c. coupling can be employed on that system as shown in Fig. 2.

The polarity of the video waveform and the variation of level with signal input at the grid of the video amplifier are now similar on both systems, hence a completely switchless video amplifier and a.g.c. system are practicable.

When this circuit is used in a complete receiver, however, there is a tendency for the receiver to 'block' on the 625-line (negative modulation) system. This phenomenon is due to the time-constant of the a.g.c. system and the overload characteristic of the i.f. amplifier. When the receiver is switched to a strong signal (that is, when the system switch is operated) the i.f. amplifier becomes overloaded before the a.g.c. control potential can build up. Under these conditions the detector output is a negative d.c. potential only, and it is blocked by the a.c. coupling to the video amplifier. The d.c. restorer diode is only capable of responding to the a.c. component of the signal, hence it is ineffective.

To prevent this effect it is apparent that not only must the video amplifier input have a signal that has a similar polarity and d.c. level on both systems, but the d.c. level on overload must also be maintained.

DOUBLE-DIODE DETECTOR CIRCUIT

The normal input to the video amplifier on the 405-line system is a positive-going video signal, and with increasing signal to the receiver the d.c. level increases positively (Fig. 1a). The use of a second detector diode, reversed in polarity from the first, enables a d.c. signal that is positive-going to be available when the normal detector produces a negative one with a negative modulation input signal. The basic circuit is shown in Fig. 3.

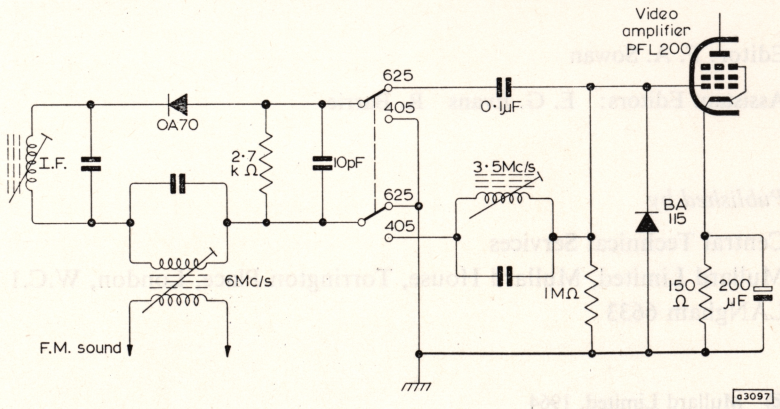

Fig. 2—Dual-standard video detector with d.c. restorer diode. The 'L' (output) section of the PFL200 is used

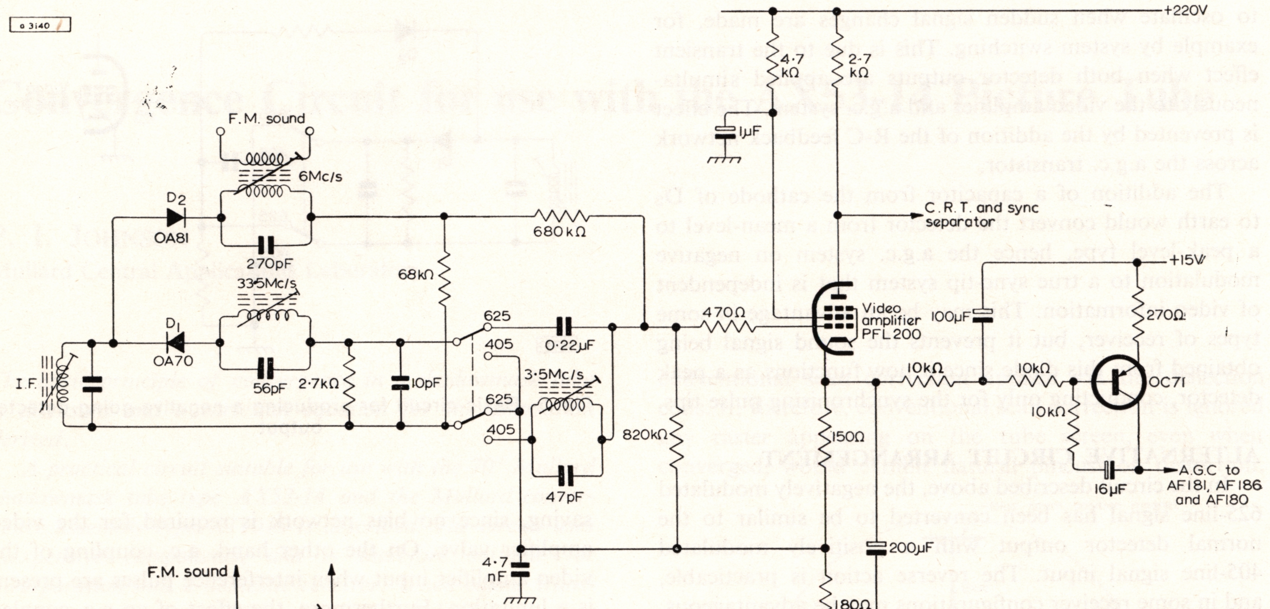

Fig. 4—Complete video detector and associated a.g.c. circuit. The 'L' (output) section of the PFL200 is used

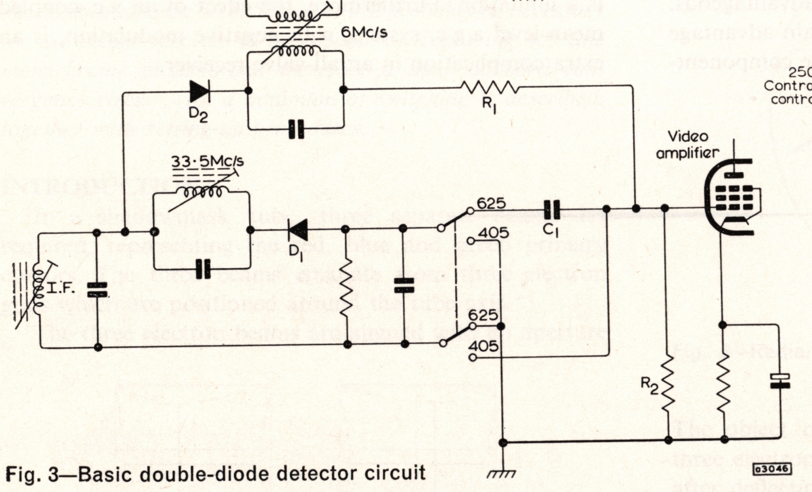

Fig. 3—Basic double-diode detector circuit

On 405 lines (positive modulation) the video amplifier is driven with the normal detector output signal from D_1. On 625 lines (negative modulation) the video signal is a.c. coupled to the video amplifier via C_1. The second video detector D_2 produces a positive-going signal, the mean of which is applied to the video amplifier, R1-C1 forming a filter circuit. The video amplifier input has the same polarity on both standards and the same d.c. level change with signal input; and, under conditions of overload, the input is positive-going. Furthermore, the 6Mc/s intercarrier sound signal may be obtained from the second detector D_2 as shown. This enables a trap circuit at the sound i.f. frequency (33·5Mc/s) to be used prior to the normal detector D_1. This circuit prevents beat frequencies between the intercarrier sound signal (6Mc/s) and video information (in particular any 4·4Mc/s colour subcarrier) from being produced and applied to the video amplifier.

HYBRID RECEIVER CIRCUIT

The complete detector circuit, video amplifier, and part of the a.g.c. system for a dual-standard hybrid receiver are shown in Fig. 4. The operation of the detector circuit is as described above, but the 33·5Mc/s i.f. trap is connected after the diode D_1 to minimise the capacitive load on the final i.f. transformer secondary. The a.g.c. amplifier transistor is driven from the cathode of the video amplifier, the a.g.c. system holding the mean cathode current constant. Manual adjustment of contrast is achieved by variation of the d.c. cathode potential which sets the d.c. input to the transistor. As the signal input to the receiver increases, the detector output on either system is positive-going, which causes the mean current of the video amplifier to increase. This increase in cathode potential tends to turn off the a.g.c. amplifier transistor, and the a.g.c. line goes negative. The receiver gain is decreased by forward a.g.c. action.

The manual contrast control is common to both systems, its range of operation being substantially the same on each. The d.c. input to the video amplifier from the second video detector is arranged to give the same video signal level on both systems by the choice of the resistor values for the resistive load network (resistors 680kΩ and 820kΩ).

At certain contrast control settings with particular signal input levels, there is a tendency for the a.g.c. system to oscillate when sudden signal changes are made, for example by system switching. This is due to the transient effect when both detector outputs are applied simultaneously to the video amplifier and a.g.c. system. The effect is prevented by the addition of the R-C feedback network across the a.g.c. transistor.

The addition of a capacitor from the cathode of D$_2$ to earth would convert the detector from a mean-level to a peak-level type, hence the a.g.c. system on negative modulation to a true sync-tip system that is independent of video information. This may be an advantage in some types of receiver, but it prevents the sound signal being obtained from this diode since it now functions as a peak detector, conducting only for the synchronising pulse tips.

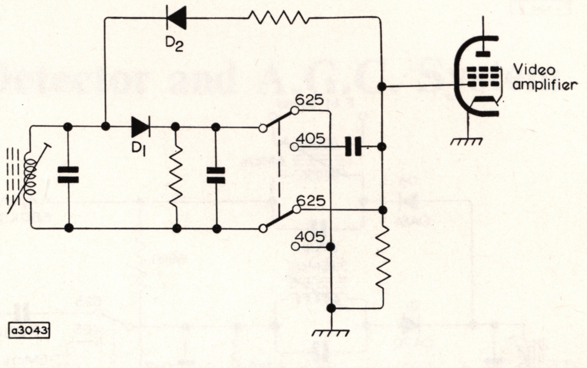

Fig. 5—Basic circuit for producing a negative-going detector output

ALTERNATIVE CIRCUIT ARRANGEMENT

In the circuit described above, the negatively modulated 625-line signal has been converted to be similar to the normal detector output with a positively modulated 405-line signal input. The reverse action is practicable, and in some receiver configurations may be advantageous. The basic circuit is shown in Fig. 5. The main advantage of this arrangement would appear to be some component-saving, since no bias network is required for the video amplifier valve. On the other hand, a.c. coupling of the video amplifier input when interference pulses are present is a limitation. Furthermore, the effect of an a.c.-coupled mean-level a.g.c. system, with negative modulation, is an extra complication in an all-valve receiver.

Convergence Circuit for use with the AX53-14 Picture Tube

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The basic principle of convergence in a shadowmask tube is discussed and a cartesian equation for the aberration derived.

A practical circuit suitable for use with the 70° Mullard shadowmask tube type AX53-14 and the Mullard convergence assembly type AT1015 is described. The circuit uses inexpensive components, and is therefore suitable for incorporation into a domestic receiver. When colour transmissions commence on 625 lines, it will probably be necessary for the colour receiver to be capable of displaying 405-line monochrome pictures and therefore a dual standard convergence circuit with a minimum of switching is described, together with setting-up procedures.

INTRODUCTION

In a shadowmask tube, three separate beams are required, representing the red, blue and green primary colours. The three beams emanate from three electron guns which are positioned around the tube axis.

The three electron beams are aligned with an aperture on the metal shadowmask so that each beam strikes the appropriate coloured phosphor dot only. This arrangement results in three rasters being produced, which overlap one another. Hence three separate pincushion distortions occur as shown in Fig. 1.

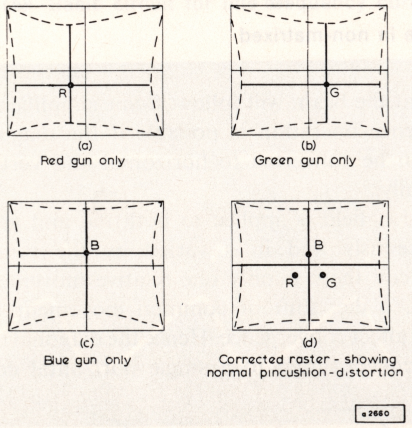

Fig. 1 (a) (b) (c)—Forms of pincushion distortion resulting from separate beams
(d)—Common pincushion distortion after convergence

CORRECTION WAVEFORMS

To correct these distortions, complex correction fields must be applied to each beam prior to deflection. Convergence is a pre-deflection correction and it in no way affects conventional scan correction applied via the deflection coils. If, therefore, conventional scan correction is ignored any raster appearing on the tube screen, even when converged, would exhibit natural pincushion distortion.

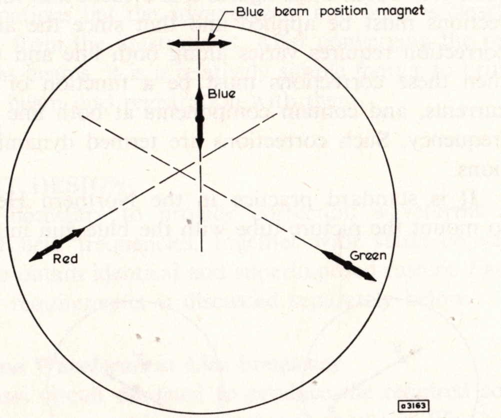

Fig. 2—Radial spot movement resulting from convergence fields

The object of convergence is solely to pre-correct the three electron beams so that the three rasters produced after deflection are identical and superimposed.

Since there are three separate electron beams in the shadowmask tube originating from physically different positions these beams (when undeflected) cannot be expected in a general case to converge on the same point on the screen. Fig. 2 shows a more typical spot configuration on the screen in such a case. It can be seen from this figure that to converge the red and green beams, suitable radial d.c. fields must be applied. Fig. 3 shows in diagrammatic form how such fields can be applied with a

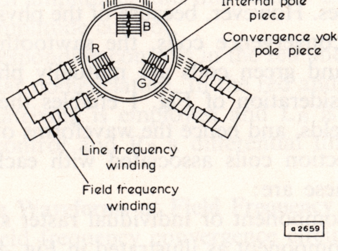

Fig. 3—Basic arrangement of the convergence yoke showing the only possible correction to be radial

typical convergence assembly. However, to bring the blue beam into convergence with the other two beams both radial and lateral components of motion are required. The radial component is achieved in the same manner as in the red and green cases but the lateral shift must be applied separately. The latter is achieved in practice by applying a field to the blue beam via a suitable pole piece (the blue beam position magnet) situated directly behind the radial correction pole pieces.

It is becoming common practice, at least in domestic receivers, to use permanent magnets in place of d.c. electromagnetic fields to provide all the static fields associated with convergence. If a complete raster is considered, after completing the d.c. (or static) convergence near the centre of the screen further errors become apparent. Figs. 1a, b and c illustrate the approximate theoretical raster shape for each gun. From these figures it is evident that further corrections must be applied and that since the amount of correction requires varies along both line and field scan then these corrections must be a function of the scan currents, and contain components at both line and field frequency. Such corrections are termed dynamic corrections.

It is standard practice in the Northern Hemisphere to mount the picture tube with the blue gun in a vertical plane so that the three guns appear at the corners of an equilateral triangle, with the blue gun uppermost. In the manufacture of the tube, compensation is made, as far as possible, for the Earth's magnetic field, for this orientation. In the Southern Hemisphere the orientation is reversed. Since the blue gun is mounted so that it is along the vertical tube axis (Fig. 1d), no convergence correction is necessary in a horizontal direction. Correction, however, is required in the vertical direction.

The red and green guns are offset from both axes and therefore require correction in both the horizontal and vertical planes. However, because of the physical positioning of the convergence coils, the sawtooth components in the red and green coils are mutually phase reversed. Further consideration of Fig. 1 enables the form of the correction fields, and hence the waveforms of the currents in the correction coils associated with each gun, to be deduced. These are:

(i) d.c. component or individual raster shift,

(ii) a.c. component as illustrated in Fig. 4.

The so-called 'parabolic' component is only approximately parabolic. A more exact form of the waveform required is shown in Appendix 1 to be a complex conic which does, however, approximate closely to a parabola over the range considered for 70° deflection.

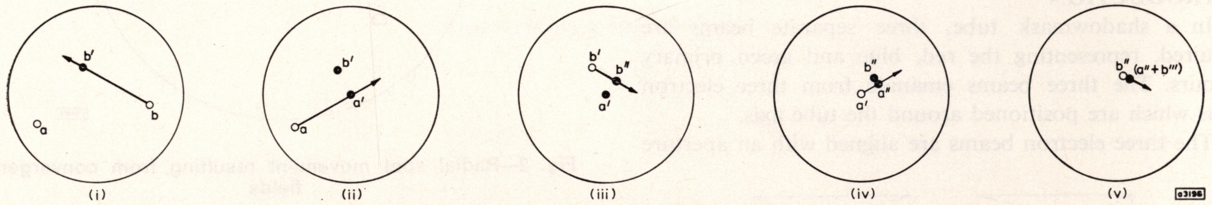

Fig. 5—Typical sequence of convergence in non-matrixed arrangement

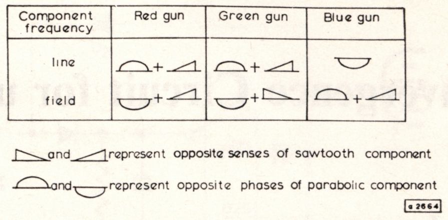
Component frequency
Red gun
Green gun
Blue gun
line
field
and
represent opposite senses of sawtooth component
and
represent opposite phases of parabolic component
a2664

Fig. 4—Correction current waveforms used for convergence

MATRIXING

A basic method of achieving convergence would be to apply the correction fields separately to each beam via an appropriate coil assembly. Fig. 3 illustrates the way in which the correction fields are applied to the electron beams. These fields can cause only a radial movement of the electron beam. If, for simplicity, a d.c. field is applied to each beam, the blue beam moves in a vertical plane but the red and green beams follow a diagonal path (Fig. 2).

Each red and green beam movement then has components of motion in both the horizontal and vertical directions such that the red and green beams follow separate diagonal paths for any single adjustment. These diagonal movements make the convergence operation difficult and laborious since it is hard to estimate the track that the beam will follow. Some simplification in the convergence operation is possible if the diagonal movement can be converted to horizontal and vertical movements only.

If a d.c. field is applied to both red and green beams simultaneously and it is varied by the same amount in each coil there is only one relative movement and, in addition, a d.c. shift or common movement. The latter does not affect convergence. Hence the diagonal movements have been converted to a single horizontal convergence correction and a common vertical shift.

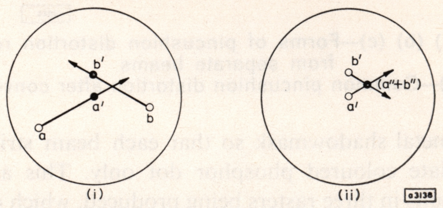

Fig. 6—Typical sequence of convergence in matrixed arrangement

This principle, a form of matrixing, can be applied to all the correction waveforms required for the red and green guns. The subjective advantage of a matrixed arrangement can be assessed from Figs. 5 and 6.

Fig. 5 shows a typical sequence of adjustments required to coverge two corresponding red and green spots on the tube screen with a non-matrixed circuit arrangement. In Fig. 5(i) spot b is moved diagonally to a position estimated to be on the locus of the other spot movement. This latter estimation is obviously subjective and since the observer can, at any one time, see only one spot move, then this estimation of the locus or 'track' of the fixed spot must inevitably be somewhat inaccurate even to some degree with a skilled observer. Similarly, spot a (Fig. 5(ii)) is now moved to a position estimated to be on the track of spot b. This procedure is then repeated as in (iii), (iv) and (v) until finally convergence is achieved – but only after a series of successive approximations.

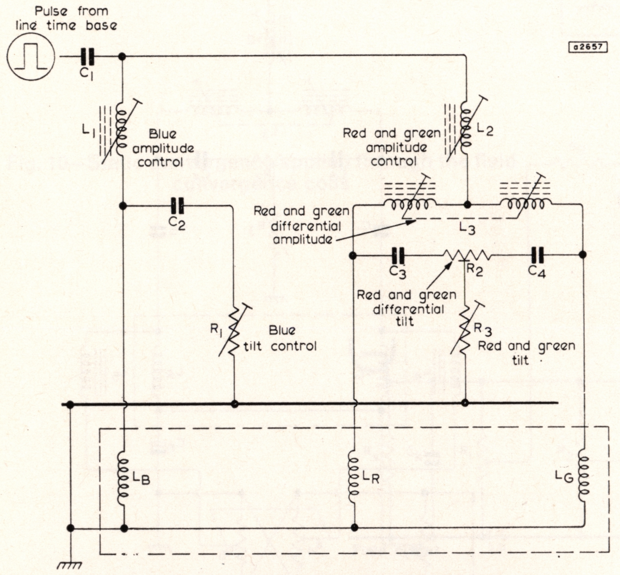

Fig. 7 (a)—Basic circuit for line frequency correction

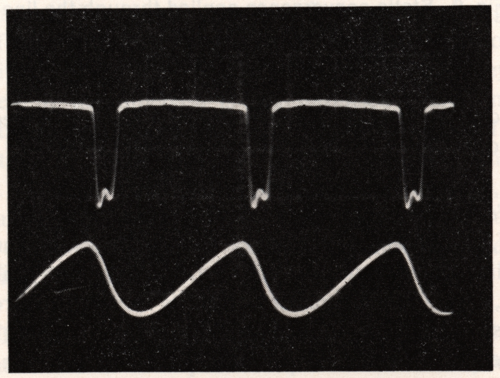

Fig. 7 (b)—Line frequency correction waveforms

It is evident that the number of these approximations depends on how accurate the 'track' estimates are. In addition, the problem is complicated by the several spot patterns necessary on the screen and by the fact that individual spot movement is not uniform over the screen area.

In a matrixed circuit arrangement, however, both electron beams move simultaneously and as shown in Fig. 6 the relative spot movement tends to be either horizontal or vertical. These simultaneous movements greatly ease the subjective problem of relative track estimation.

Fig. 6 shows a typical sequence when a matrixed circuit is used. The operation of an 'overall' control increases the fields applied to the two beams by the same percentage and both spots move as shown. Because of tube tolerances the guns cannot be supposed to be symmetrically disposed around the tube axis, and therefore an overall movement cannot alone bring about convergence. Thus a 'differential' control is required and by this means the field applied to one beam is increased at the same time as the field applied to the other is decreased. This results in the movement shown in Fig. 6(ii).

The sequences of operation shown in Figs. 5 and 6 are idealised to illustrate the principles of convergence. In practice, both methods would require a series of successive approximations but the advantage of matrixing is clearly apparent from the illustrations. After converging the red and green beams, it is a relatively simple matter to bring the blue beam into registration with them.

CIRCUIT DESIGN

It is necessary to provide correction waveforms at line and field frequencies, together with static convergence, to obtain identical and superimposed rasters. Each of these requirements is discussed separately below.

Correction Waveforms at Line Frequency

A basic circuit designed to produce the required correction waveforms at line frequency is shown in Fig. 7(a). For simplicity the blue beam section of the circuit is considered first.

A pulse from the line transformer of approximately 80V amplitude is integrated by L_1, producing a sawtooth current waveform which is fed into the circuit comprising L_B, C_2, and R_1. This circuit is tuned to approximately two-thirds of the line frequency and since it has a relatively low Q, the circulating current takes the form of a line frequency sinewave with second harmonic distortion. This waveform is a very close approximation to a parabola. As R_1 is increased the sawtooth component of current in L_B becomes more prominent, moving the minimum of the parabola to the left, as shown in Fig. 7(b). This control is therefore termed a tilt control. The adjustment of L_1 provides an overall adjustment of the parabolic and sawtooth amplitudes and this control is consequently called the amplitude control.

The corresponding controls in the red and green circuit have a similar nomenclature. In the red and green circuit, however, matrixing is employed and L_B is a differential amplitude control and R_2 a differential tilt control.

Convergence Waveforms at Field Frequency

In the field frequency convergence circuit (Fig. 8) the two component waveforms are derived separately. The sawtooth component is taken directly from a tertiary winding on the field output transformer. This tertiary winding has a centre tap connected to earth, and provides both negative and positive going sawtooth waveforms.

Opposite senses of sawtooth component of current are required for the red and green coils while at the same time the required parabolic component is in the same phase (see Fig. 4). Consequently, it is not possible to use a simple matrix circuit.

If a resistor is connected from one side of the tertiary winding to one side of the convergence coils (R_{15}, Fig. 8), opposite senses of sawtooth waveform can be produced across R_{11}. Adjustment of R_{11} varies the amplitude of both waveforms simultaneously, since with movement in one direction the waveforms tend to cancel. Adjustment of R_9 causes a differential variation by a similar process. Hence R_{11} functions as the overall tilt control and R_9 as the differential tilt control.

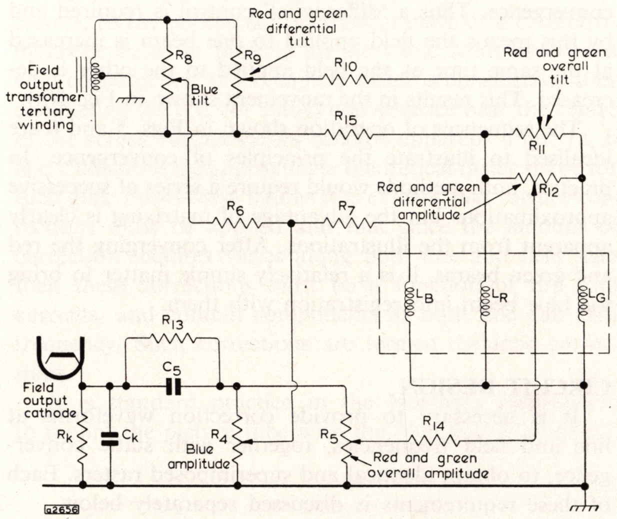
Field output transformer tertiary winding
R8
Blue tilt
R9
Red and green differential tilt
R10
Red and green overall tilt
R15
R11
R12
Red and green differential amplitude
R6
R7
LB
LR
LG
R13
C5
Field output cathode
Rk
Ck
R4
Blue amplitude
R5
R14
Red and green overall amplitude
a2656

Fig. 8—Field frequency convergence circuit

The parabolic component is derived from the cathode circuit of the field output valve. The bypassed cathode resistor effectively integrates the sawtooth current so producing the required parabola. Matrixing is again employed in the red and green circuits with R_5 acting as the overall amplitude control and R_{12} the differential amplitude control.

Capacitor C_5 is used solely for d.c. blocking to reduce the necessity of altering the static convergence after each field circuit adjustment. It is shunted by R_{13} to insert some compensating d.c. component. R_6 and R_7 are included as stand-off resistors to reduce the interdependence of R_8 and R_4. Resistors R_{10} and R_{14} are included for a similar purpose.

Static Convergence

Static convergence is achieved by steady magnetic d.c. fields. Such fields can be derived by passing a direct current through the line or field convergence coils, or small permanent magnets can be employed. Any overall picture shift, however, must be provided for via the main deflection coils.

Static Convergence in the Line Frequency Circuit

Fig. 9 illustrates the principle of providing static convergence by passing suitable direct currents through the line convergence coils. These latter coils are, however, of relatively low inductance and to provide a satisfactory correction a current of approximately 200mA could be required. In Fig. 9 the basic convergence circuit is unaltered and the only additional components are extra potentiometers R_1, R_4 and R_5 together with suitable stand-off inductors L_4 and L_5. To be effective, these inductors must be of relatively low resistance and approximately 250mH inductance. Such components would be considered uneconomical in a commercial receiver design.

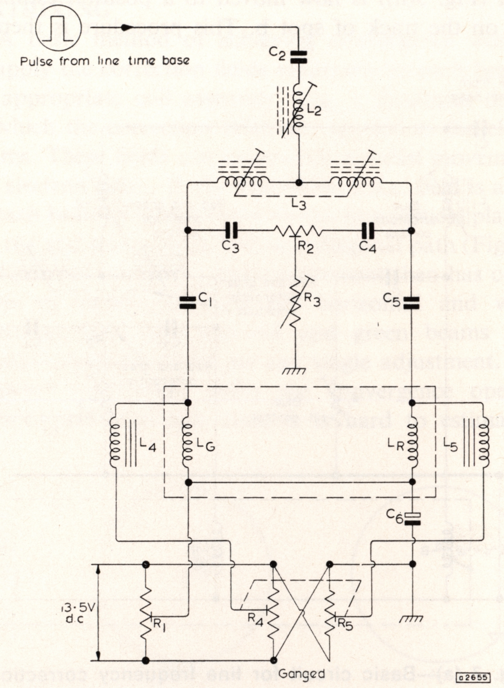

Fig. 9—Static convergence applied through the line convergence coils

Static Convergence in the Field Frequency Circuit

If, instead of passing the d.c. correction current through the line convergence coils, the field frequency coils are used, a very much smaller current is required to produce the same correction field, because of the much larger inductance of the field frequency coils. In practice, it is possible to use the d.c. potential at the cathode of the field output valve, and a basic circuit is shown in Fig. 10.

As in the line frequency circuit, the a.c. network remains unaltered although isolating capacitors C_1, C_4, C_5 and C_6 are incorporated. To reduce interaction of the d.c. convergence with the a.c. corrections, filters R_9, C_7, R_{10}, R_7, C_3, R_8 are also incorporated. Additional components such as electrolytic capacitors and resistors are required in this type of circuit and the current of the field output valve must be stabilised to prevent variation of the static convergence with mains input variations and valve ageing. Hence this circuit is rather uneconomical for a commercial receiver.

ADDITIONAL FACTORS AFFECTING CONVERGENCE

There are additional external factors which can affect convergence. These include mains supply, temperature and e.h.t. variations. It is also necessary to allow for the effect of external magnetic fields which can result in purity errors. Therefore, precautions against extraneous causes of misconvergence are discussed separately below.

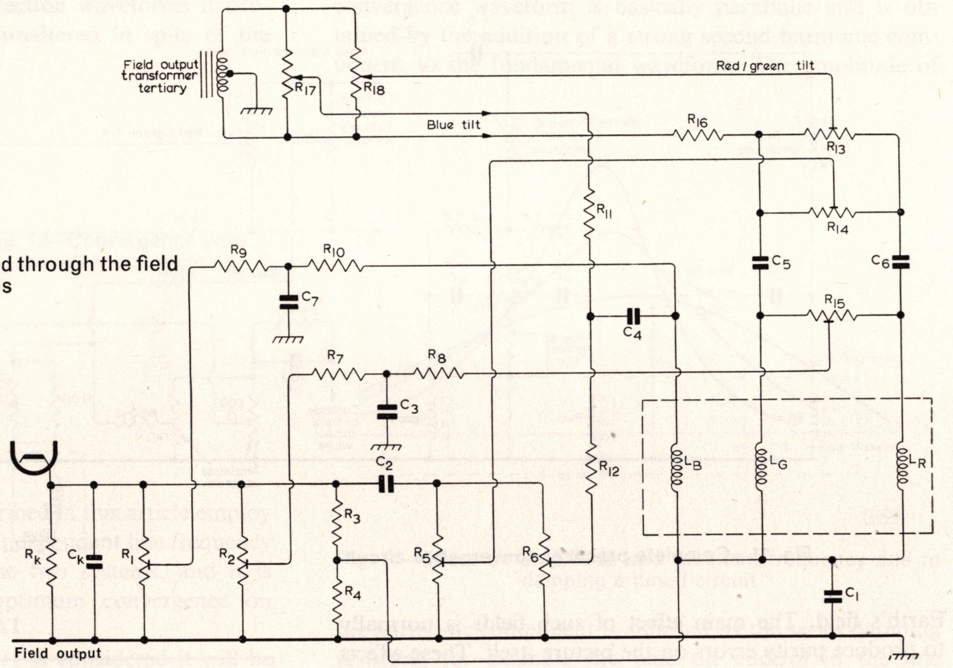

Fig. 10—Static convergence applied through the field convergence coils

Mains Supply Variations

The amplitudes of the line frequency convergence waveforms are, naturally, directly proportional to the amplitude of the line flyback pulse from which they are derived. If the timebase is not stabilised against mains variations, some convergence errors can be expected with variations in supply potential. Supply variations can also affect the waveforms derived from the field timebase, with similar results. However, it is now almost standard practice to provide some form of scan stabilisation for the line and field timebases.

E.H.T. Variations

Variations in e.h.t. potential produce a velocity variation in the electron beam which in turn produces a variation in deflection as the beam passes through any given field. Therefore, some misconvergence does occur as the e.h.t. is varied. The degree of misconvergence for a given e.h.t. change is dependent on the correction field applied and hence will vary from tube to tube. With an average tube an e.h.t. variation of 3kV, for example, corresponding to a beam current change of 1mA and source impedance of 3MΩ, results in only minor convergence errors at the tube centre. The effect of any such misconvergence can be minimised if during the convergence procedure the tube beam current is kept at approximately one-half of its maximum value, for example, $400\mu A$. In this way the errors in any one direction are halved.

The most useful convergence test pattern is a white grid or spot pattern and the mean picture tube beam current has, therefore, a very low value. The increase in e.h.t. current to some $400\mu A$ during the convergence operation is therefore most simply achieved by the use of an e.h.t. shunt resistor of some 60MΩ.

Timebase Linearity

The shapes of the field frequency convergence waveforms depend directly on the linearity of the field output stage from which they are derived. Consequently, it is always good practice to adjust the linearity of both timebases before attempting a convergence procedure.

Purity Adjustments

Any alteration in the purity correction field, or even the position of the main scan coils, can profoundly affect convergence. It is, therefore, essential that all purity adjustments are made before attempting any convergence procedure.

Temperature Compensation

To minimise convergence changes with ambient temperature variations it is essential that the circuits are compensated. To assess the extent of this variation, some temperature measurements were carried out. The main temperature dependent factor was the resistance of the field convergence coil, which rose by 17% for a temperature increase of 50degC. This variation in resistance can be compensated for by the use of a suitable temperature dependent resistor. The Mullard thermistor VA1034, shunted by a 120Ω resistor, connected in series with each coil provided satisfactory compensation.

A change in field timebase linearity can arise when the resistance of the field deflection coils increases with temperature. This change of linearity could result in some misconvergence. However, it is common practice to incorporate a thermistor in the field deflection coils circuit, to maintain constant amplitude and linearity with temperature variations.

Magnetic Fields

Any shadowmask receiver is very sensitive to the effect of external magnetic fields, even those of the order of the Earth's field. The main effect of such fields is normally to produce purity errors on the picture itself. These effects may become apparent after the receiver has been physically moved from one part of a room to another or from one room to another, or, in some cases, after merely rotating it through 90°. Any small errors which occur in this way can normally be corrected by adjustment of the purity magnets but if this is done then it will probably be found that some re-convergence is required.

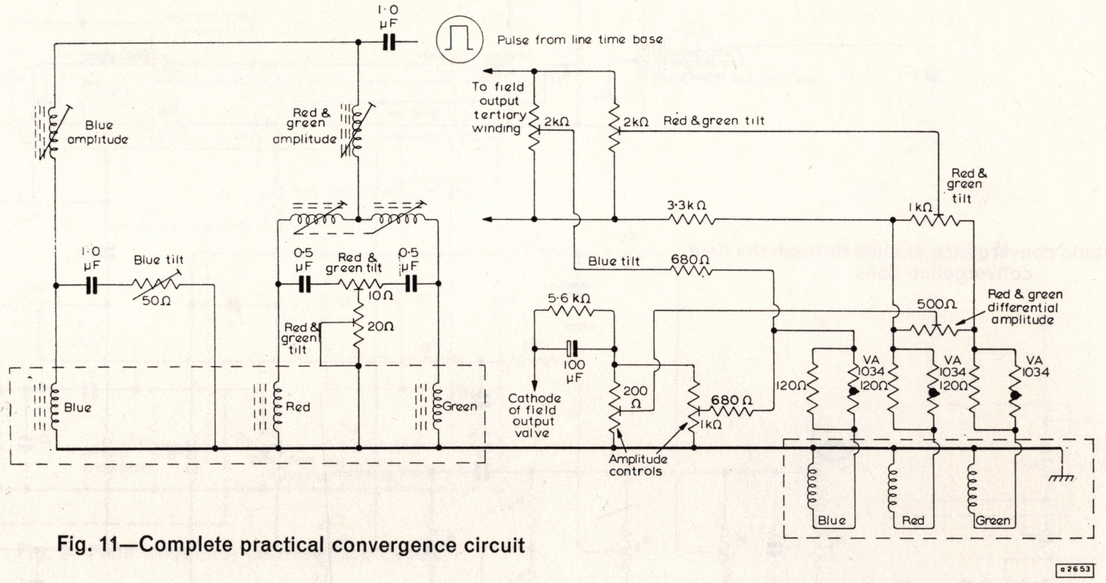

Fig. 11—Complete practical convergence circuit

The most satisfactory solution is to degauss the whole receiver, that is, subject it to a large alternating field, before the tube is originally set up and after the set is in its final position. If this procedure is carried out, then any purity errors which subsequently occur can be corrected by further degaussing without upsetting the convergence. In a receiver a simple magnetic shield is necessary around the flare of the tube to enable the user to move the receiver without the necessity for degaussing.

PRACTICAL RECEIVER CIRCUIT

The complete convergence circuit developed for use in an experimental receiver is shown in Fig. 11. In this receiver the convergence assembly is a Mullard type AT1015. The assembly connections are shown in Fig. 12.

A positive 80V pulse from the line output transformer is applied to the line convergence circuit via a 1μF capacitor (Fig. 11). The field output transformer tertiary winding provides 24V peak-to-peak drive waveform (centre tap earthed) and a parabolic waveform of 7V peak-to-peak is derived from the cathode circuit of the field output valve.

Setting up procedures for purity and convergence are given in Appendices 2 and 3. The parabolic correction waveform is an approximation of the required waveform (see Appendix 1) and therefore perfect convergence will not be obtained around the edges of the tube, the error being typically 2mm.

CONVERGENCE COIL DETAILS

Details of the convergence coils are given in Table 1.

TABLE 1

Red and Green Amplitude Coil
250 turns 35 s.w.g. enss Cu $\frac{1}{2}$in wide
Douglas Winding Gears:
48 32 31 46 24 96 ($\frac{1}{4}$ wave)
L = 300μH to 2·5mH R = 4·5Ω

Red and Green Differential Amplitude Coil
2×300 turns 35 s.w.g. enss Cu 0·3in spacing $\frac{1}{2}$in wide
L = 0·4 to 3·5mH R = 5Ω
Winding gears as Red and Green Amplitude Coil

Blue Amplitude Coil
300 turns 29 s.w.g. enss Cu $\frac{1}{2}$in wide
Douglas Winding Gears:
39 34 32 36 24 96 ($\frac{1}{4}$ wave)
L = 600μH to 3·6mH R = 2·5Ω

All these experimental coils were wound on Aladdin 'long' formers with Ferroxcube cores (Mullard FX1054) mounted on 0BA brass studding. Neosid ferrite screw cores have been found satisfactory in a laboratory receiver, but they reduce the inductance swing by approximately 25% (Neosid 6×1×25mm F14).

DUAL STANDARD CONVERGENCE

Colour programmes will commence, in the United Kingdom, on the 625-line standard as an addition to the normal monochrome programmes. Until complete programme duplication occurs, however, it will probably be necessary for a colour receiver to operate satisfactorily when displaying a 405-line monochrome picture.

Convergence errors are more noticeable on a monochrome display, and it is therefore essential that the convergence in a dual standard receiver should be of a high standard on both systems.

Circuit Requirements

The basic requirement of any dual standard convergence circuit is that the correction waveforms it produces shall remain completely unaltered in spite of the system change. The circuits described in this article employ duplicate and therefore entirely independent line frequency convergence components for the two systems, and it is therefore possible to attain optimum convergence on both standards.

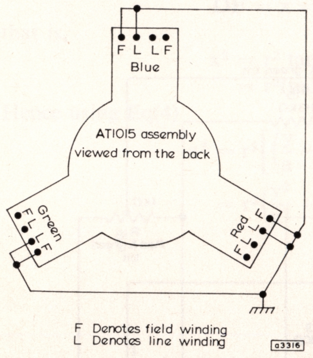

Fig. 12—Convergence yoke assembly connections

If the circuit shown in Fig. 11 is considered it will be apparent that the circuit consists of two sections:

(i) field frequency section,
(ii) line frequency section.

As the field frequency is common to both standards, the waveforms produced by the field frequency section are unaffected by any system change.

The waveforms derived at line frequency, however, are inherently frequency dependent. The parabolic component of these waveforms is derived by means of a frequency selective circuit consisting of the actual convergence yoke inductance and a suitable tuning capacitor, the circuit being resonant at approximately two-thirds of the line frequency. The waveform amplitude is controlled by means of a variable inductance, and is therefore also frequency dependent. Hence, if convergence is to be maintained on both standards, every variable in the line frequency section must be switched with the line frequency.

The frequency selective circuit is of relatively low Q, and the tuning is, therefore, not critical. The required convergence waveform is basically parabolic and is obtained by the addition of a strong second harmonic component to the fundamental waveform. The amplitude of the harmonic component is adjusted by the series damping resistors, for example the blue tilt control in the line circuit of Fig. 11. The variation in the ratio of second harmonic component to the fundamental as the damping resistor is adjusted is illustrated in Fig. 13.

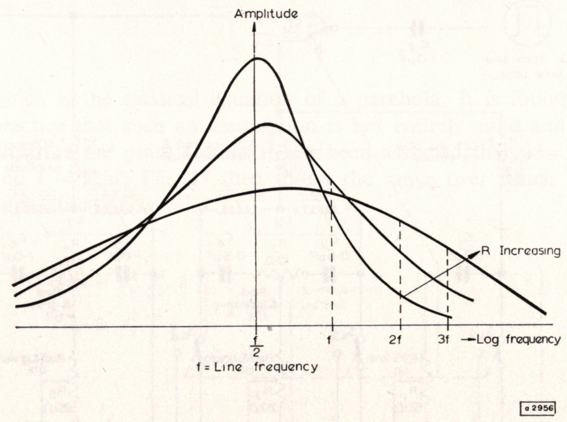

Fig. 13—Effect on amplitude and resonant frequency due to damping a tuned circuit

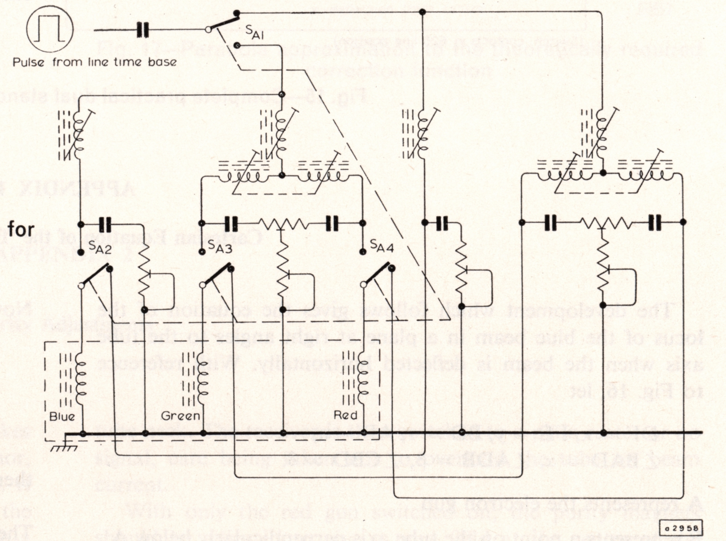

Fig. 14—Duplicate circuit with simple switching operation for dual standard working

In choosing the value of the tuning capacitor it is necessary to ensure that resonance occurs with negligible third harmonic component present. It has been found that a resonant frequency of about two-thirds the line frequency is satisfactory.

Switching Requirements

If all the frequency dependent components in the line frequency circuits have to be modified for each standard it is evident that the switching requirements and the resulting setting-up procedure would be somewhat difficult. A far simpler arrangement can be achieved, however, if instead of modifying the one circuit to be dual standard, a duplicate circuit is provided. The arrangement in Fig. 14 is practicable because of the relatively simple nature of the basic circuit. The actual switching can be achieved by means of a 4-pole 2-way switch. With such a system, two sets of line frequency controls are provided, but it has been found that once convergence has been achieved on one standard the adjustment of the duplicate controls on the other standard is a relatively simple operation.

In a dual standard receiver, despite the switching modifications, the circuit performance and the supply voltage requirements remain unchanged from the requirements given for the 625-line convergence system. The complete dual standard convergence circuit is shown in Fig. 15.

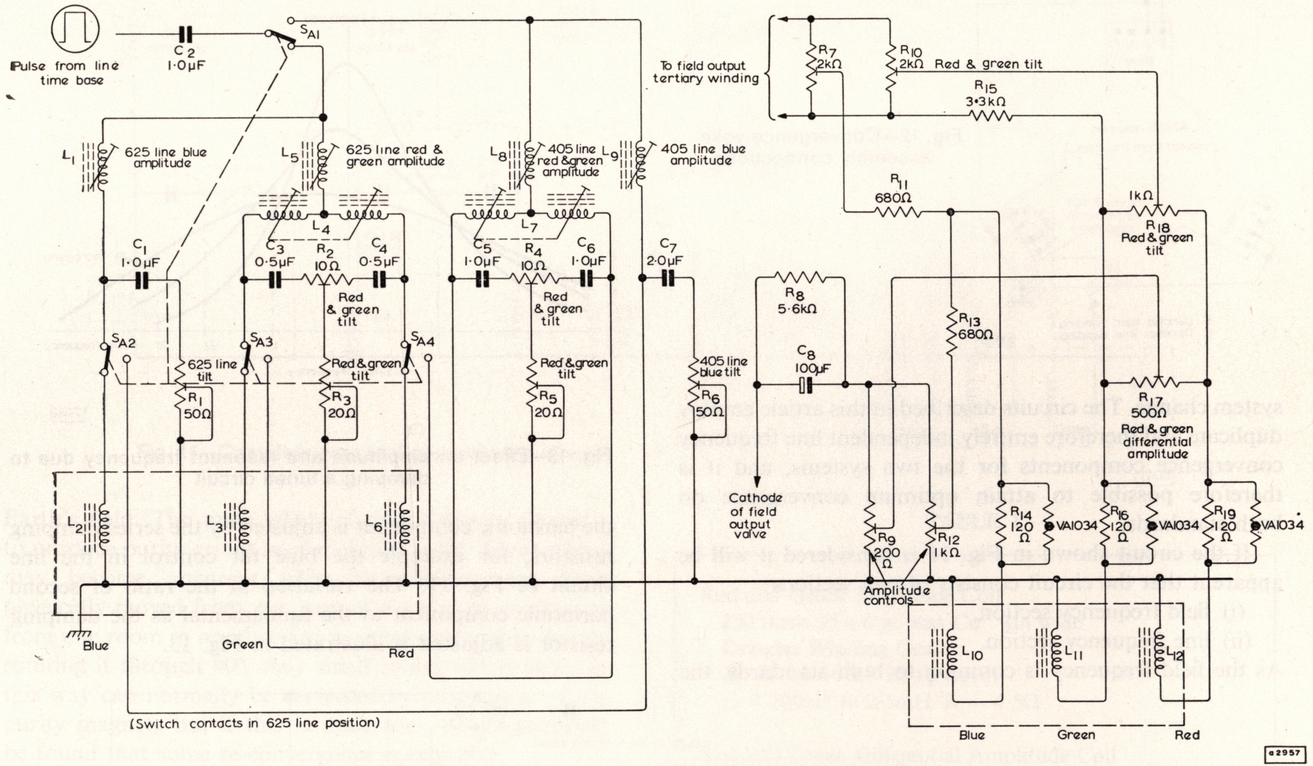
Pulse from line time base
C2 1·0µF
SA1
L1 625 line blue amplitude
L5 625 line red & green amplitude
L8 405 line red & green amplitude
L9 405 line blue amplitude
L4
L7
C1 1·0µF
C3 0·5µF
R2 10Ω
C4 0·5µF
C5 1·0µF
R4 10Ω
C6 1·0µF
C7 2·0µF
Red & green tilt
SA2
SA3
SA4
625 line tilt
R1 50Ω
Red & green tilt
R3 20Ω
Red & green tilt
R5 20Ω
405 line blue tilt
R6 50Ω
L2
L3
L6
Blue
Green
Red
(Switch contacts in 625 line position)
To field output tertiary winding
R7 2kΩ
R10 2kΩ
Red & green tilt
R15 3·3kΩ
R11 680Ω
1kΩ
R18 Red & green tilt
R8 5·6kΩ
C8 100µF
R13 680Ω
R17 500Ω
Red & green differential amplitude
Cathode of field output valve
R9 200 Ω
R12 1kΩ
Amplitude controls
R14 120 Ω
VA1034
R16 120 Ω
VA1034
R19 120 Ω
VA1034
L10
L11
L12
Blue
Green
Red
a2957

Fig. 15—Complete practical dual standard convergence circuit

APPENDIX 1

Cartesian Equation of the 'Blue Parabola'

The development which follows gives the equation of the locus of the blue beam in a plane at right angles to the tube axis when the beam is deflected horizontally. With reference to Fig. 16, let

$$DF = x,\ FE = y,\ BD = r,\ AB = a,$$
$$\angle BAD = \alpha,\ \angle ADB = \beta,\ \angle CBD = \delta.$$

A represents the electron gun

B represents a point on the tube axis perpendicularly below A

$\angle CAD$ represents the angle of deflection

DFE represents a plane at right angles to the tube axis

AD represents the electron beam which is deflected horizontally through an angle CAD to position AE. This beam cuts the horizontal plane through BD at C at an angle β, that is, points C and D are on the arc of a circle of radius r and centre B.

Now

$$BC = BD = r$$
$$AC = AD = r \operatorname{cosec} \alpha$$
$$\angle CFE = 90^\circ$$
$$\angle ACB = \angle FCE$$

therefore,

triangles ABC and FCE are similar.

Therefore,

$$\frac{CF}{BC} = \frac{EC}{AC} = \frac{FE}{AB} = \frac{y}{a}\,. \qquad \ldots(1)$$

But

$$\angle ACB = \angle ADB = \beta\,.$$

Therefore,

$$\tan \beta = \frac{y}{CF} = \frac{a}{r}$$

that is,

$$y = \frac{a}{r}.CF\,. \qquad \ldots(2)$$

Considering triangle BDF

$$\cos \delta = \frac{r}{r+CF}$$

therefore,

$$CF = r(\sec \delta - 1)\,. \qquad \ldots(3)$$

From Eq(2)

$$y = r(\sec \delta - 1)\frac{a}{r}$$
$$= a(\sec \delta - 1)\,. \qquad \ldots(4)$$

Also from triangle BDF

$$DF = x = r \tan \delta$$

that is,

$$x^2 = r^2 \tan^2 \delta$$
$$= r^2(\sec^2 \delta - 1)\,.$$

Hence using Eq(4)

$$x^2 = r^2\left[\left(\frac{y}{a} + 1\right)^2 - 1\right]$$
$$= r^2\left(\frac{y^2}{a^2} + \frac{2y}{a} + 1 - 1\right)$$

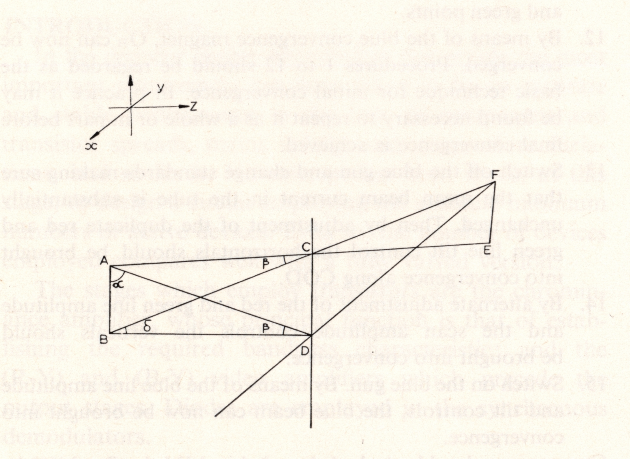
y
z
x
A
B
C
D
E
F
α
β
δ

Fig. 16—Convergence geometry for one beam

therefore,

$$\frac{x^2}{r^2} - \frac{y^2}{a^2} = \frac{2y}{a}\,. \qquad \ldots(5)$$

This is the equation of a complex conic. The equation as so derived is a function of distance but if the scanning waveform is assumed to be a linear function of time, that is

$$y = \phi(t)\,,$$

then Eq(5) can also be considered as a function of time.

It is perhaps of interest to note that if the y^2/a^2 term is ignored then the locus becomes

$$\frac{x^2}{r^2} = \frac{2y}{a}$$

which is the classical equation of a parabola. It is found in practice that such an assumption is not entirely valid and to illustrate the point further it has been assumed that a = ¼in and r = 12in; Fig. 17 then shows the range over which the approximation is valid.

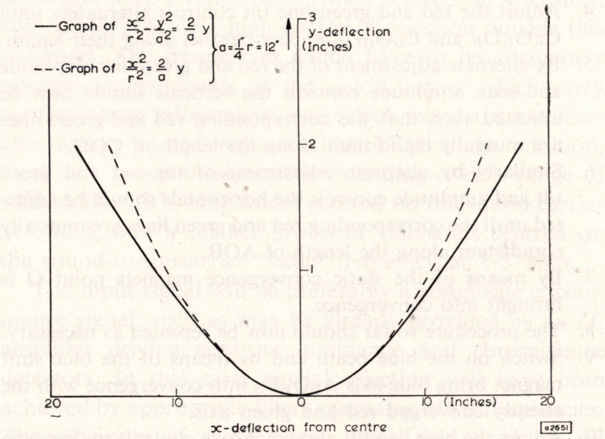

Fig. 17—Parabola approximation to the theoretically required correction function

APPENDIX 2

Purity Adjustments

If the beam from any one electron gun after deflection strikes the phosphor corresponding to that gun and only that phosphor, then the colour produced on the screen is said to be pure. It is the object of the following purity adjustments to correct the beams prior to convergence so that all three guns produce a pure raster over the whole face of the tube.

First, when the receiver is placed in its operating position it must be degaussed, that is, it is subjected to a large alternating field which is gradually reduced to zero, to remove extraneous magnetic fields. Then the purity magnets on the back of the convergence yoke are set at minimum (two tabs together) and the main scan coils moved back as far as possible along the tube neck. The receiver is then switched to a plain raster or no signal, care being taken not to overload the tube on beam current.

With only the red gun switched on, the purity magnets should be adjusted to give a pure field at the centre of the screen and also to cover the greatest possible screen area. With the red field pure the other two fields should also be satisfactory but it may be found necessary to readjust the magnets on the green field and so reach a state of compromise.

Purity assessment can be very difficult and critical examination of the screen is necessary if extremely good results are to be achieved.

APPENDIX 3

Convergence Setting-up Procedure

Before any convergence can be attempted the whole receiver should be degaussed in the position in which it is desired that it be operated and the purity adjusted in the normal way. Both timebases should be adjusted for optimum linearity.

The following procedure should then be followed, bearing in mind throughout that the theoretical best a convergence circuit of this type can do is to converge the tube about its axes, that is along AOB and COD in Fig. 18 at all points along their length.

1. Switch the receiver to a white grating pattern and set all the convergence controls to their mid position. (If no shunt regulator valve is employed, the e.h.t. load current should be increased to about 400μA by an external load resistor.)
2. Adjust the static convergence magnets, together with the blue shift magnet, such that O_R and O_G appear about $\frac{1}{8}$in apart with O_B between them.
3. Switch off the blue beam. Adjust the red and green field tilt and amplitude controls alternately until $A_GO_GB_G$ and $A_RO_RB_R$ are parallel.
4. Adjust the red and green line tilt controls alternately until $C_RO_RD_R$ and $C_GO_GD_G$ appear parallel along their length.
5. By alternate adjustment of the red and green line amplitude and scan amplitude controls the verticals should now be adjusted such that the corresponding red and green lines are mutually equidistant along the length of COD.
6. Similarly by alternate adjustment of the red and green tilt and amplitude controls the horizontals should be adjusted until the corresponding red and green lines are mutually equidistant along the length of AOB.
7. By means of the static convergence magnets point O is brought into convergence.
8. The procedure so far should now be repeated as necessary.
9. Switch on the blue beam and by means of the blue shift magnet bring blue axis $A_BO_BB_B$ into convergence with the already converged red and green axis.
10. Adjust the blue line tilt and amplitude controls to converge at COD.

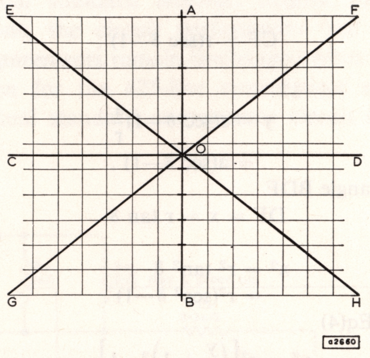

Fig. 18—Convergence axes

11. The horizontal crossing points along AOB should be adjusted to be mutually equidistant from the relative red and green points.
12. By means of the blue convergence magnet, O_B can now be converged. Procedures 1 to 12 should be regarded as the basic technique for initial convergence. In practice it may be found necessary to repeat it as a whole or in part before final convergence is achieved.
13. Switch off the blue gun and change standards making sure that the mean beam current in the tube is substantially unchanged. Then by adjustment of the duplicate red and green line tilt control the horizontals should be brought into convergence along COD.
14. By alternate adjustment of the red and green line amplitude and the scan amplitude controls the verticals should be brought into convergence.
15. Switch on the blue gun. By means of the blue line amplitude and tilt controls, the blue beam can now be brought into convergence.

Convergence should now have been achieved on both standards.

Hybrid Decoder for N.T.S.C. Equiband Colour Receivers

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The decoder described, designed for N.T.S.C. U.K. standards, accepts a low-level chrominance signal and a reference signal in the (R-Y) or (B-Y) phase. High-level colour difference outputs are directly connected to the grids of a shadowmask picture tube. Readily available transistors are used in the low-level stages, and valves in the three high-level output stages.

INTRODUCTION

In the decoder design described in this article the most important requirement has been taken to be an accurate and stable performance in the face of component and transistor spreads, mains voltage and temperature variations, and so forth. No attempt has been made, at this stage, to design a circuit employing the absolute minimum numbers of active devices. Even so, the number of devices employed compares well with a fully valved version.

The stages which employ transistors are the chrominance amplifier, whose principal function is that of establishing the required bandpass characteristic; and the (R-Y) and (B-Y) video amplifiers which precede the output stages. Diodes are employed in the synchronous demodulators.

With the addition of a subcarrier regenerator unit, the decoder is suitable for use in a domestic receiver.

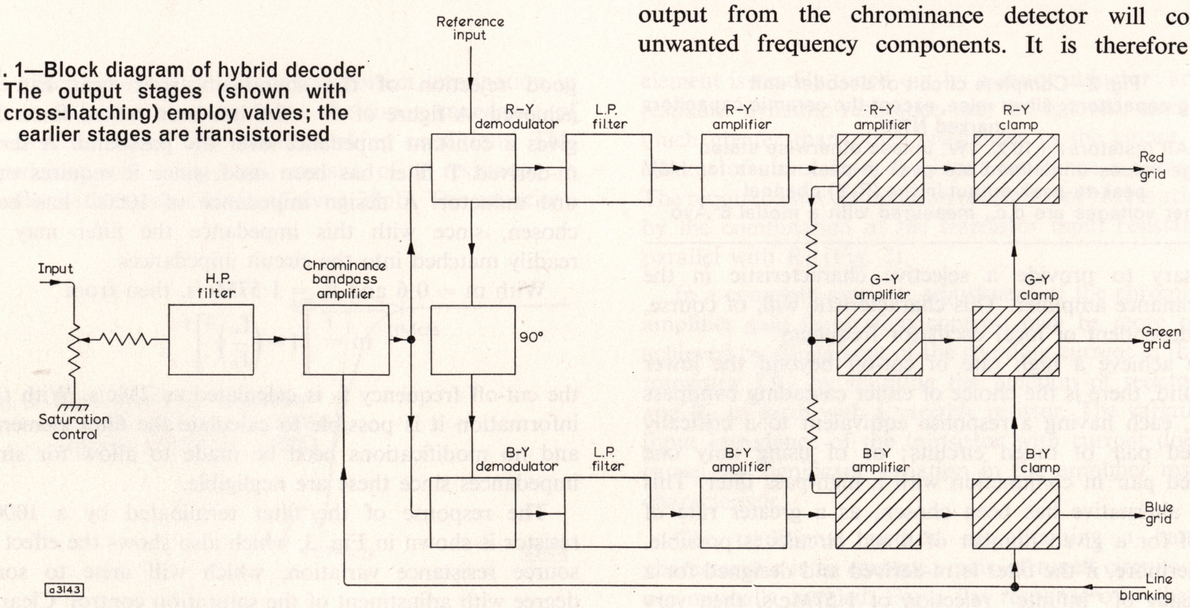


Fig. 1—Block diagram of hybrid decoder
The output stages (shown with cross-hatching) employ valves; the earlier stages are transistorised

CIRCUIT DESIGN

A block diagram, Fig. 1, shows the general form of the unit; while Fig. 2 shows the complete circuit diagram. The individual "blocks" will now be considered in detail.

Chrominance Bandpass Amplifier

In an equiband receiver, the chrominance bandpass amplifier has to accept sidebands of the subcarrier without distortion over a range of ±1Mc/s centred on 4·43Mc/s, and has to provide a high degree of rejection outside this passband. The frequencies which have to be adequately rejected are sidebands of the vision carrier which carry the luminance information, and which may have considerable energy in the region below 3·4Mc/s. In addition, there is a group of sidebands extending over a range of 150kc/s (for 10% amplitude) centred on the sound carrier at 6Mc/s, and a further group of sidebands centred on the sound-to-subcarrier "beat" frequency at 1·57Mc/s.

The input signal will be preferably a composite chrominance signal, such as may be already provided by an i.f. amplifier driving separate luminance and chrominance detectors. In this case, partial selection will have been achieved by appropriate filters operating at i.f. frequencies. The efficiency of these filters, since they work at i.f. frequencies, will naturally be dependent upon the local oscillator frequency. It follows that under conditions of maladjustment of fine-tuning, or oscillator drift, the output from the chrominance detector will contain unwanted frequency components. It is therefore still

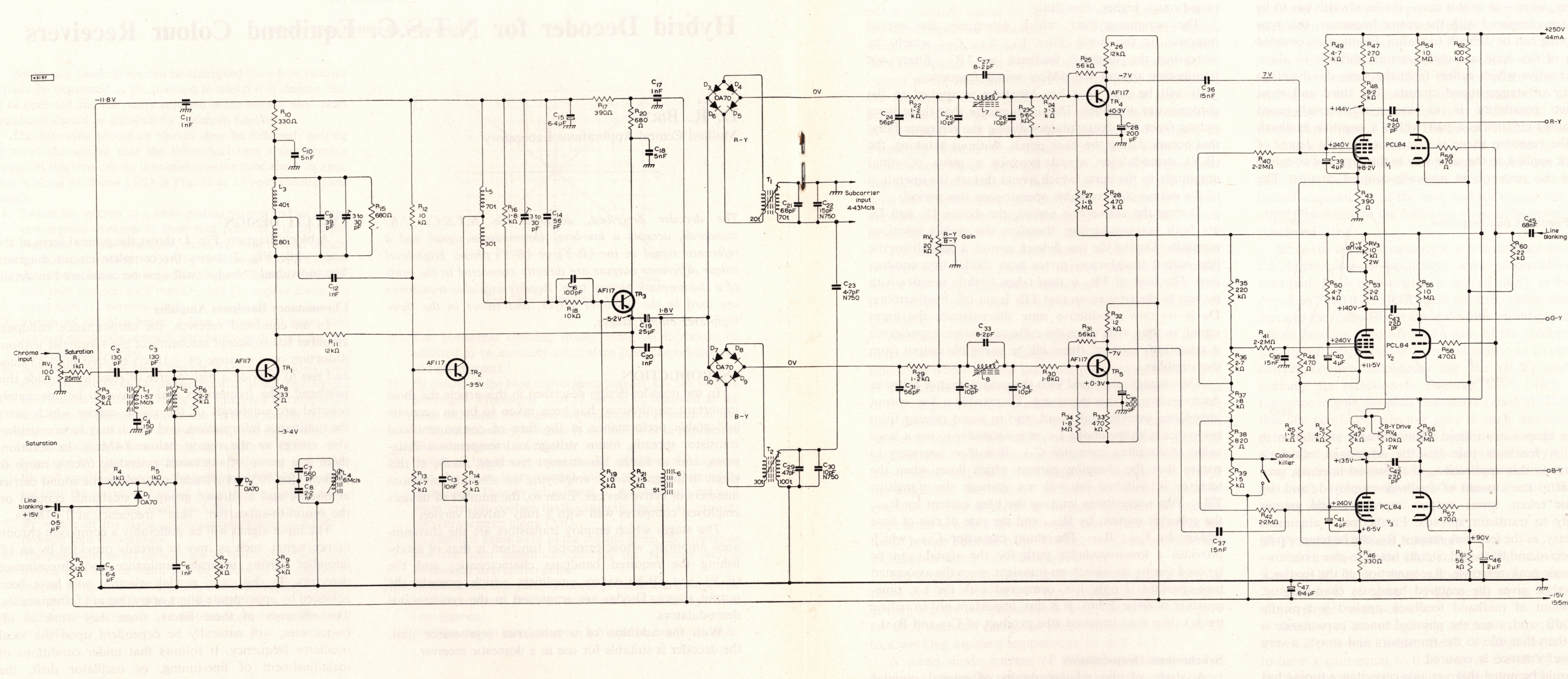


Fig. 2—Complete circuit of decoder unit
Tuning capacitors: silver mica, except the ceramic capacitors marked N750
All resistors: ±10%, ½W, unless otherwise stated
Voltage values underlined are peak-to-peak values for 140V peak-to-peak output in the (R-Y) channel
Other voltages are d.c., measured with a model 8 Avo

necessary to provide a selective characteristic in the chrominance amplifier. This characteristic will, of course, be independent of local oscillator frequency.

To achieve a high rate of cut-off beyond the lower sideband, there is the choice of either cascading bandpass filters, each having a response equivalent to a critically coupled pair of tuned circuits; or of using only one coupled pair in conjunction with a high-pass filter. This latter alternative has been chosen, as a greater rate of cut-off for a given number of tuned circuits is possible. Furthermore, if the filter is m-derived and designed for a frequency of "infinite" rejection of 1·57Mc/s, then very good rejection of the sound-subcarrier beat can be achieved. A figure of m = 0·6 has been taken, since this gives a constant impedance over the passband. A series m-derived T filter has been used, since it requires only one inductor. A design impedance of 1000Ω has been chosen, since with this impedance the filter may be readily matched into the circuit impedances.

With m = 0·6 and f_∞ = 1·57Mc/s, then from

$$m = \left[1-\left(\frac{f_\infty}{f_c}\right)^2\right]^{\frac{1}{2}}$$

the cut-off frequency f_c is calculated as 2Mc/s. With this information it is possible to calculate the filter elements, and no modifications need be made to allow for stray impedances since these are negligible.

The response of the filter terminated by a 1000Ω resistor is shown in Fig. 3, which also shows the effect of source resistance variation, which will arise to some degree with adjustment of the saturation control. Clearly, this effect is negligible. To ensure correct termination in circuit, it is necessary to know the input impedance of the first transistor TR_1 under working conditions. This impedance will, in general, consist of a resistance in parallel with a capacitance of some 25pF. The capacitive element is readily tuned out by a shunt inductor; and the resultant dynamic resistance may be ignored, since it is much greater than other resistances in the circuit. At a current of about 2mA the input resistance is some 2kΩ. The required filter termination resistance of 1kΩ is achieved by the combination of the transistor input resistance in parallel with R_6 (Fig. 2).

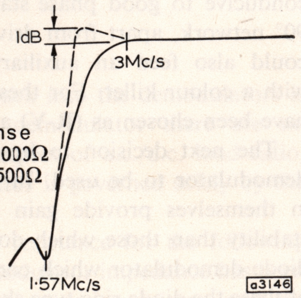
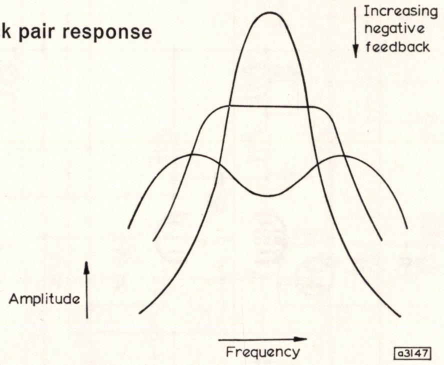


Fig. 3—High-pass filter response
——— source resistance = 1000Ω
– – – source resistance = 500Ω

In a complete receiver, adjustment of the chrominance amplifier gain (saturation control) may be more readily achieved by adjustment of the standing current in the first transistor TR_1 by shunting the junction of resistors R_3 and R_4 to earth with a variable resistor. The variation in input impedance of the transistor with current does not cause any significant variation in the amplifier response characteristic.

The traditional method of obtaining a bandpass characteristic is to employ a pair of tuned circuits which are mutually-coupled. For a flat response over the passband, pure mutual inductance coupling is to be preferred. However, when—as in this case—the bandwidth has to be very large compared with the centre frequency, this type of coupling can be difficult to realise. In any case, coupled circuits of this kind are comparatively difficult to align. An alternative which suffers from the same disadvantage is a pair of stagger-tuned circuits. The third and most attractive possibility is to use synchronously-tuned single-tuned circuits as a pair within a negative feedback loop. The response shape is determined by the degree of feedback applied, in the same way as the degree of coupling governs the response in mutually-coupled circuits. The response shapes are indeed identical, and are shown in Fig. 4. A feedback pair has the advantage, which is important in this application, of improved linearity, proportional to the amount of feedback employed; and, by the same token, lower output impedance and greater immunity to transistor spreads. Furthermore, alignment is very easy, as the feedback resistor R_{11} can be temporarily disconnected and the tuned circuits tuned to give a narrow-band single-peak response. Reconnection of the feedback resistor then gives the required bandpass characteristic. The amount of midband feedback applied is typically some 10dB; and, since the physical tuning capacitance is greater than that due to the transistors and strays, a very stable performance is ensured.

Fig. 4—Feedback pair response



It should be noted that variable capacitance tuning has been used, and that no slugs are employed in L_3 and L_5. This has been done to avoid the inherent change in effective turns ratio which takes place when tuning is accomplished by permeability tuning. In most cases it is possible to change transistors in the stage TR_2 without any noticeable change in gain or response shape taking place, and without the need for re-tuning. The loss of gain, in comparison with mutually-coupled circuits, is only some 4dB, which is no embarrassment in this case.

The need, in the chrominance amplifier, for a low output impedance over the band 50c/s to 5·5Mc/s arises because the two demodulators are fed in parallel from this same point. Clearly, any common impedance at this point, of a magnitude that is significant in relation to the load, will upset the isolation between the two channels (R-Y) and (B-Y). The load on the circuit from the demodulators is some 700Ω, and this cannot easily be raised, since it is determined by impedances in the output circuits of the demodulators. Because of practical difficulties in filter design, and the loading imposed by the following transistor stages, the impedance of the output circuits cannot be raised much higher than 2kΩ.

The remaining filter, which determines the overall response, is the notch filter L_4, C_7, C_8, which, by increasing the negative feedback to TR_1, filters out frequencies around the 6Mc/s sound frequency.

It will be seen that line blanking is applied to the chrominance amplifier. This is to ensure that there is no output from the demodulators during the reference burst that occurs during the back porch. Without blanking, the (B-Y) demodulator would produce a pulse of equal amplitude to the burst, which would disturb the operation of the output clamps which operate over this period.

During the active line period, the diodes D_1 and D_2 are both non-conducting, therefore the circuit functions normally. During the line flyback period, a pulse from the line output transformer drives both diodes into conduction. The base of TR_1 is thus taken slightly positive with respect to the emitter, so that TR_1 is cut off. Furthermore, D_2 in its now conductive state short-circuits the input signal, so that, even when the collector-to-base capacitance is taken into account, there will be negligible output from the amplifier.

The circuits associated with the output emitter-follower deserve some further comment. To provide a low output impedance over the full band, and to avoid ringing from resonance with the choke L_6, it is necessary to use a large value of coupling capacitor C_{19}. It is then necessary to ensure that the charging current which flows when the receiver is switched on will not damage the transistor TR_3. This necessitates limiting the base current by R_{18}, the collector current by R_{20}, and the rate of rise of base voltage by C_{15}, R_{17}. The shunt capacitor C_{16}, which provides a low-impedance path for the signal, can be ignored during the switch-on transient, since the associated time-constant is only 1μs, compared with the h.t. time-constant of some 2·5ms. It is thus important not to reduce the h.t. line time constant (the product of C_{15} and R_{17}).

Synchronous Demodulators

A study of the relative merits of several popular demodulator-matrix configurations (Ref. 1) has revealed that operation along the (R-Y) and (B-Y) axis followed by matrixing to obtain (G-Y) showed considerable advantages over other systems when the effects of component tolerances and spreads in demodulator gains and detection phases were taken into account. Also, it is only necessary to provide one phase-shift network of 90° outside the automatic phase control loop. This in itself is conducive to good phase stability. It is possible that this 90° network, apart from driving the (B-Y) demodulator, could also feed an auxiliary phase-detector associated with a colour killer. For these reasons the detection axes have been chosen as (R-Y) and (B-Y).

The next decision to be taken concerns the type of demodulator to be used. In general, types which do not in themselves provide gain offer a greater promise of stability than those which do. There are several types of diode demodulator which come in the first category, and of these the diode ring type shows a number of advantages. The use of this diode configuration is well known as a means of generating a double sideband suppressed carrier signal, but it is equally useful for detecting such a signal. One of the most important advantages is that the reference signal, which is large in comparison with the chrominance signal, does not appear at the input to the demodulator or at the output point, because of the bridge balance. In addition, the amplitude of the reference signal is not important, so long as it is large enough to control the diode switching at all times. This is quite an important advantage, as frequently there is some disturbance of the amplitude of the reference signal provided by the subcarrier regenerator during the burst interval, due to operation of the phase-detector. Furthermore, the amplitude of reference signal applied to each demodulator need not be identical, which eases the design of the phase-shift network.

The impedances are such that a wide bandwidth can be provided for the detected output signal in this case, where the highest output frequency (1Mc/s) is not very far removed from the lowest sideband input frequency (3·4Mc/s).

Although the reference signal is not present in the output, there will be components of the chrominance sidebands which are for the most part concentrated close to the subcarrier frequency of 4·4Mc/s. Some filtering of the output signal is therefore still required. It is found that this filter is quite able to cope with some residual reference signal which will be present if, because of diode spreads, the bridge is not perfectly balanced. The subsequent video amplifiers have also been designed to accept the associated d.c. output. In addition, the chrominance amplifier is not affected by unbalance, since its output impedance is only a few ohms. In view of the fact that the diode bridge balance is not critical, no extra resistors are included in series with the diodes.

The required 90° phase shift is provided by a pair of tuned circuits coupled by "top" capacitance. A measure of temperature compensation is used, and a typical figure for the maximum change in phase shift from cold (20°C) to a working ambient temperature of 38°C is 2°.

A mean diode current of 0·5mA, measured as d.c. flowing in the reference winding, is necessary for correct functioning of the demodulators. In the design described this is achieved with an input of about 10V peak-to-peak. The diode conduction angle is close to 180°, and with this wide conduction angle there should be no trouble from radiation of harmonics of the subcarrier.

The decoder will operate with a reference signal either in the (R-Y) phase, as shown in Fig. 2, or with a signal in the (B-Y) phase applied to the (B-Y) demodulator.

Low-pass Filters

The impedance looking back into the demodulators is some 1000Ω. With a ring demodulator there are no particular requirements in respect of a.c./d.c. load ratio, as the operation of the detector is governed solely by the reference signal.

Since a convenient input resistance for the following video stage is 2·2kΩ, the demodulator impedance has been padded out to this value with a series resistor, and the filter design impedance taken as 2·2kΩ.

An m-derived low-pass filter has again been used in preference to a notch type, as it provides attenuation for harmonics of the subcarrier frequency which are present to an appreciable extent.

As was the case for the high-pass filter at the input of the chrominance amplifier, the procedure has been to choose m = 0·6 for constant impedance, and to place the frequency of high attenuation f_∞ at 4·43Mc/s in the centre of the chrominance band. The cut-off frequency can then be determined, and comes out to be 3·55Mc/s, which is satisfactory. The filter component values can then be calculated, but some allowance has now to be made for the self-capacitance of the coil, and for circuit strays. The input impedance to the following stage is, however, almost purely resistive.

Since the cut-off frequency is a safe distance away from the highest signal frequency, some mismatch can be tolerated. Such a mismatch is indeed likely to be introduced by the (B-Y)/(R-Y) gain control, which shunts the input to the filter. This is a pseudo-differential gain control, whose function is to allow the gain of the channel which has highest amplification to be reduced so as to obtain the correct relative amplitudes for the (R-Y) and (B-Y) outputs. Its inclusion is essential if a ±10% overall tolerance in gain for each channel is aimed at. This means that the combined effective gain of each combination of demodulator, video amplifier, and output stage must be controlled to within ±1dB. Even though every effort has been made towards stability, this specification cannot be met with 10% components without some differential pre-set adjustment.

(R-Y) and (B-Y) Video Amplifiers

The function of the (R-Y) and (B-Y) video amplifiers is to amplify the low-level outputs from the demodulators up to a level of some ±3·5V peak, so as to drive the output stages satisfactorily.

Because of the weighting applied at the transmitter to avoid over-modulation, the (R-Y) amplifier is required to have a gain equal to 0·56 that of the (B-Y) amplifier. Further, the gain of these amplifiers must be stable, as differential changes will result in errors in colour reproduction. The input impedance is required to be 2·2kΩ resistive, and the bandwidth at least a little greater than 1Mc/s to allow for bandwidth shrinkage when cascaded with other restricted-bandwidth stages.

In addition, linearity must be good. This means that the operating point must not shift with changes in temperature, as the signal amplitude is quite large compared with the h.t. supply of 15V. Furthermore, to avoid the need for an input coupling capacitor, operation should be undisturbed by a residual d.c. output from the demodulators which may occur through diode unbalance.

Though this is quite a formidable list of requirements, they can all be met quite easily in practice. Apart from a minor difference in the input network designed to provide the required gain ratio, both stages are identical. Since a +250V supply is available in the receiver, this is made use of in conjunction with a very large emitter resistor of 470kΩ to give very good d.c. stability. A further feedback path, effective at d.c. as well as at signal frequencies, is inserted between the collector output point and the base input (R_{25}). The current flowing in this collector-to-base resistor is balanced by current bled from the +250V line through a 1·8MΩ resistor. Allowance is made for the base current. Thus, even at the extremes of the normal spreads in current gain, the current flowing in the signal source (demodulator) is insufficient to cause degradation of performance. In short, the configuration resembles that of a collector follower with a high degree of d.c. current feedback. However, the feedback components are dimensioned to provide a gain of some 20 times for the (B-Y) amplifier. The signal feedback under these conditions is typically 15dB.

This form of circuit will present a load to the demodulators which has an a.c./d.c. ratio less than unity. However, as mentioned earlier, this does not introduce any difficulty with a ring demodulator.

Colour Difference Output Stages and Matrix

The colour difference output stages employ valves. This is because the output voltage swing may be of the order of 150V peak-to-peak, whilst the working bandwidth is 1Mc/s. At present, this requirement is more easily met with valves than with transistors.

The circuit configuration employed is substantially that previously described (Ref. 1). It will be seen that all three amplifiers employ negative feedback, and that portions of the signal developed across the (R-Y) and (B-Y) cathode feedback resistors are added to give the (G-Y) signal at the cathode of the (G-Y) grounded-grid amplifier. The resistor R_{47} is included as a common impedance in the anode circuits of the (R-Y) and (B-Y) amplifiers to eliminate crosstalk which would otherwise arise from the common impedance in the cathode circuits.

The circuit is inherently insensitive to component spreads and valve ageing, because of the combined effects of cathode degeneration and d.c. screen grid feedback. To ensure long valve life, each valve runs at approximately half of its anode dissipation rating, and at one third of its cathode current rating. The peak current demand is considerably less than the current available from a nominal new valve, so that a considerable loss in emission can be tolerated before there is any noticeable effect.

A slight modification has been made to the output arrangements used in the original circuit (Ref. 1). Previously, the pre-set drive adjustments were made by potentiometers, in which the slider was connected to the clamp circuit and c.r.t. grid, and the track to the output valve anode. It has been found that a neater layout, with reduced stray capacitance, can be achieved if the adjustment is made by varying the anode load resistance. This is particularly so, in this case, where the clamp triode occupies the same valve envelope as the output pentode.

Provisions for a colour killer have been made in the bias arrangements, so that all three colour difference amplifiers can be switched off by an electronic switch capable of passing 1mA d.c. and withstanding a voltage of 15V nominal.

The use of a.c. coupling to the grids of the colour difference amplifiers may be expected to have some transient effect on the signal-handling capacity. The effect occurs after a fast transition from programme material involving fully saturated colour difference signals with a large d.c. component, to material where there is substantially no d.c. component (for example, colour bars). In the former case, if we consider a simple example where the d.c. component is fully retained up to the coupling capacitor, this capacitor acquires a change of charge equal to the signal d.c. component. After the transition, there will be some disturbance to the bias conditions of the colour difference amplifiers, which may result in distortion of the colour difference signals, for a period determined by the time-constant of the coupling network. The worst condition arises when the bias shift is towards cut-off, when the valve will be prevented from drawing grid current. With a coupling capacitor of 15nF and a grid leak of 2·2MΩ, such a disturbance will have become negligible in a period of two fields.

In view of the fact that this momentary error is so short-lived, and that the conditions which can give rise to such an error are most unlikely to occur with normal programme material, it is not considered that a.c. coupling has any practical disadvantage, provided that the d.c. component is reinserted at the output of the colour difference amplifiers.

It is, however, wise not to make the coupling time-constants too long; but, at the same time, it must be remembered that they must be longer than the clamp time-constant, otherwise the clamp could, under certain conditions, remain inoperative for a number of lines.

Clamp Circuits

The d.c. component of the colour difference signals is reinserted at the grids of the picture tube. By reinserting the d.c. component in this way, at the latest possible point, the highest stability is ensured.

Because colour difference signals can, and do, have either polarity, some form of keyed-clamp, operated by a pulse from the line output transformer, is necessary.

Equilibrium has to be reached during the line flyback pulse, so that the time-constant applying to the coupling capacitor clamp impedance and output stage anode load resistor, has to be short. This leads to a coupling capacitor value of some 220pF.

In order to prevent undue "sag" during the active line period, the clamp load resistor (R_{54}, for example) has to have a value of some 10MΩ. In view of this high impedance and the large voltage swing, a valve is thought to be more attractive than a semiconductor, at the present time, particularly as the triode required can be combined with the colour difference amplifier in a single envelope.

The level to which the c.r.t. grids are clamped can be varied by changing the values of R_{61} and/or R_{62}.

CONSTRUCTION

A prototype has been constructed on a chassis measuring 8½×4¾in (22×12cm) which is shown in Fig. 5. In general, the construction is not critical, and the screening employed is more elaborate than required. The main requirement is to confine the subcarrier reference signal to the areas where it is needed.

The upper section houses the chrominance amplifier, and below this are the two demodulators and the 90° phase-shift circuits. The lower section houses the low-pass filters, all the colour difference amplifiers, and the clamps. The tuned coils are all wound on Brayhead miniature formers with cans; and, with the exception of the low-pass filter coils, all employ single-layer solenoid construction. Full constructional details are given in Figs. 6 to 13.

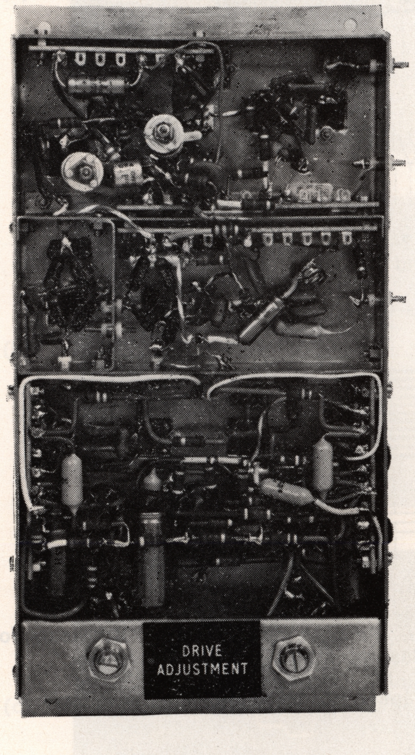


Fig. 5—Chassis layout (underside)

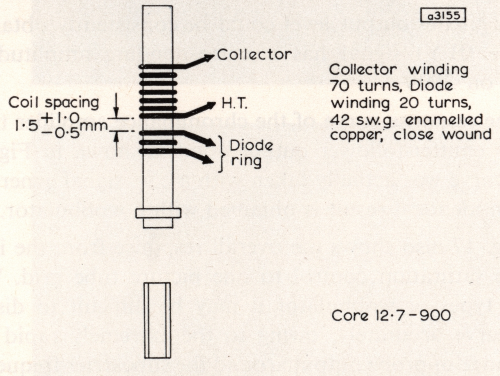

a3155
Collector
H.T.
Coil spacing
1·5 +1·0 −0·5 mm
Diode
ring
Collector winding 70 turns, Diode winding 20 turns, 42 s.w.g. enamelled copper, close wound
Core 12·7−900

Fig. 6—Winding data for T_1
Can and base as for L_1 (Fig. 8)
The core must be inserted from the lower end, so that it is positioned under the diode winding

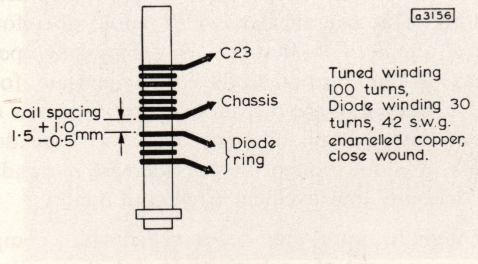

a3156
C23
Chassis
Coil spacing
1·5 +1·0 −0·5 mm
Diode
ring
Tuned winding 100 turns, Diode winding 30 turns, 42 s.w.g. enamelled copper, close wound.

Fig. 7—Winding data for T_2
Can and base as for L_1 (Fig. 8). Core as for T_1 (Fig. 6)
The core must be inserted from the lower end, so that it is positioned under the diode winding

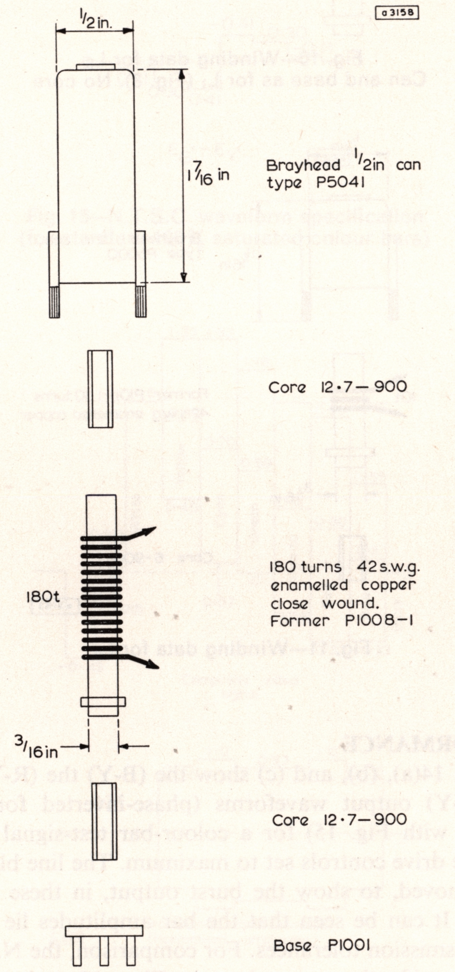

a3158
1/2in.
1 7/16 in
Brayhead 1/2in can type P5041
Core 12·7−900
180t
180 turns 42 s.w.g. enamelled copper close wound. Former P1008−1
3/16in
Core 12·7−900
Base P1001

Fig. 8—Winding data for L_1
Two cores are used. Total inductance = 66·5μH

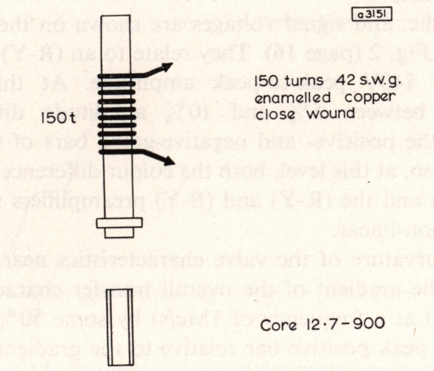


Fig. 9—Winding data for L_2
Can and base as for L_1 (Fig.8)

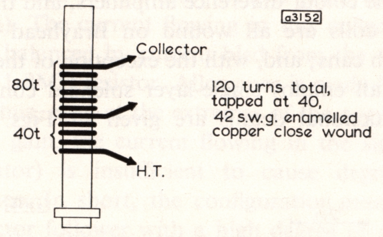

a3152
Collector
80t
40t
120 turns total, tapped at 40, 42 s.w.g. enamelled copper close wound
H.T.

Fig. 10—Winding data for L_3
Can and base as for L_1 (Fig. 8). No core

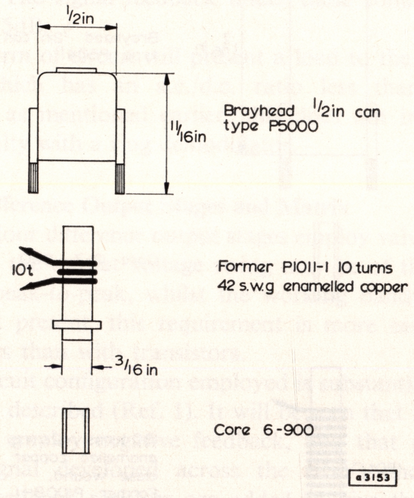

½in
Brayhead ½in can type P5000
11/16in
10t
Former P1011-1 10 turns 42 s.w.g enamelled copper
3/16in
Core 6-900
a3153

Fig. 11—Winding data for L_4

PERFORMANCE

Figs. 14(a), (b), and (c) show the (B-Y) the (R-Y) and the (G-Y) output waveforms (phase-inverted for comparison with Fig. 15) for a colour-bar test-signal input, with the drive controls set to maximum. The line blanking was removed, to show the burst output, in these photographs. It can be seen that the bar amplitudes lie within the transmission tolerances. For comparison, the N.T.S.C. signal specifications are given in Figs. 15 and 16. The time scale is 10μs/cm, and the vertical voltage scale 20V/cm. The slight overshoots visible on some bars can be seen also on the input waveform, Fig. 14(e), and are thus not generated within the unit. Fig. 14(d) shows the chrominance signal at the output from the emitter-follower TR_3.

Both d.c. and signal voltages are shown on the circuit diagram, Fig. 2 (page 16). They relate to an (R-Y) output signal of 140V peak-to-peak amplitude. At this level there is between 5% and 10% amplitude difference between the positive- and negative-going bars of the test signal. Also, at this level, both the colour difference output amplifiers and the (R-Y) and (B-Y) preamplifiers start to become non-linear.

The curvature of the valve characteristics near cut-off reduces the gradient of the overall transfer characteristic (measured at a frequency of 1Mc/s) by some 50% in the region of peak positive bar relative to the gradient at the origin.

With a 250V h.t. line, it is doubtful whether an appreciably greater output level could be consistently obtained, as the (B-Y) signal has a corresponding amplitude of close on 200V.

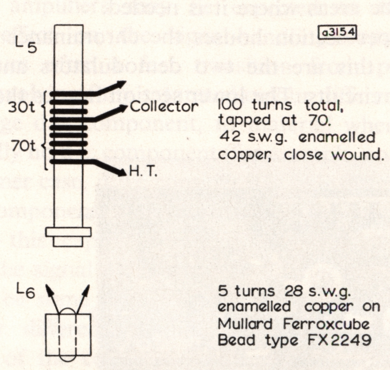

a3154
L5
30t
70t
Collector
100 turns total, tapped at 70. 42 s.w.g. enamelled copper, close wound.
H.T.
L6
5 turns 28 s.w.g. enamelled copper on Mullard Ferroxcube Bead type FX2249

Fig. 12—Winding data for L_5 and L_6
L_5 can and base as for L_1 (Fig. 8). No core

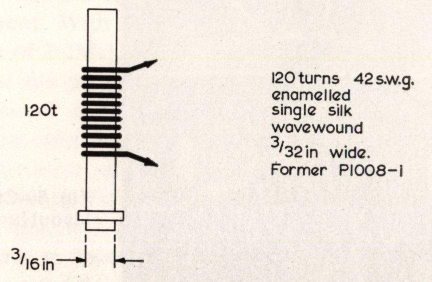

120t
120 turns 42 s.w.g. enamelled single silk wavewound 3/32in wide. Former P1008-1
3/16in

Fig. 13—Winding data for L_7 and L_8
Can, cores, former, and base as for L_1 (Fig. 8). Gears for Douglas winder: 50, 32, 44, 36, 40, 80
Inductance = 120μH
Typical self-capacitance (in can) = 5·5pF at 32Mc/s

The response curve of the chrominance amplifier taken at the emitter-follower output point is shown in Fig. 17. This curve was actually taken with a c.w. signal generator, but an identical result is obtained with a wobbulator.

Fig. 17 also shows the overall response from the input at the saturation control to one picture tube grid. With some types of wobbulator it may be difficult to display this curve accurately, owing to the extremely rapid rate of change of probe output around the subcarrier frequency. However, if a separate wideband oscilloscope is connected to the picture tube grid, the envelope can be observed quite readily. The overall bandwidth is restricted by the limited bandwidth of the high-level colour difference amplifiers; but it is comparable to the transmitter 'Q' bandwidth. The overall bandwidth could therefore be increased slightly, if desired, by employing peaking inductors in the output leads to compensate for the capacitive loading of the picture tube grids. In view of the associated introduction of cross colour and quadrature cross-talk, it is not thought that an increase in bandwidth would yield any improvement in picture quality.

Although the unit is intended to be fed with a composite chrominance signal, the selectivity is such that quite good results can be obtained with a composite video input signal.

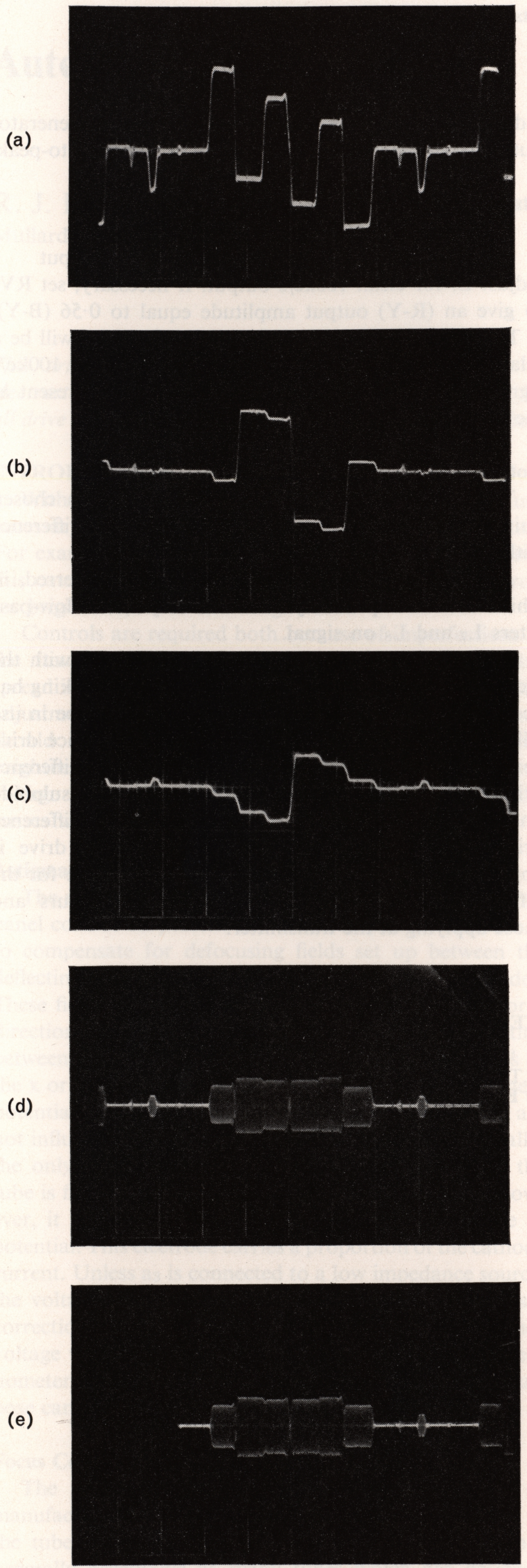


Fig. 14—Output waveforms (graticule, 1cm squares)
(a) B–Y output (b) R–Y output (c) G–Y output
(d) Chrominance signal at output from emitter-follower TR_3
(e) Input

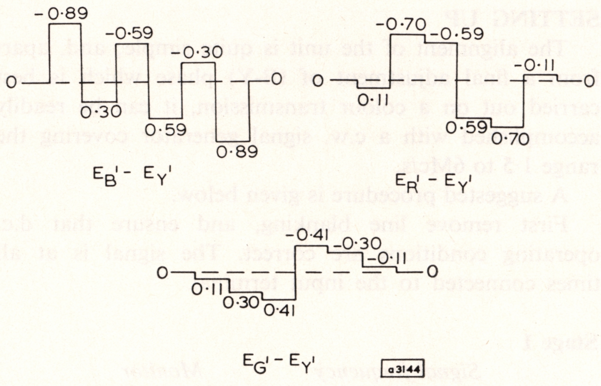


Fig. 15—N.T.S.C. waveform specification (for standard 100% saturated colour bars)

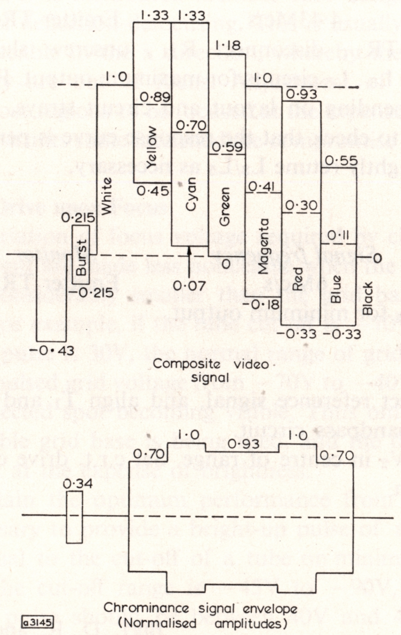


Fig. 16—N.T.S.C. colour difference waveforms

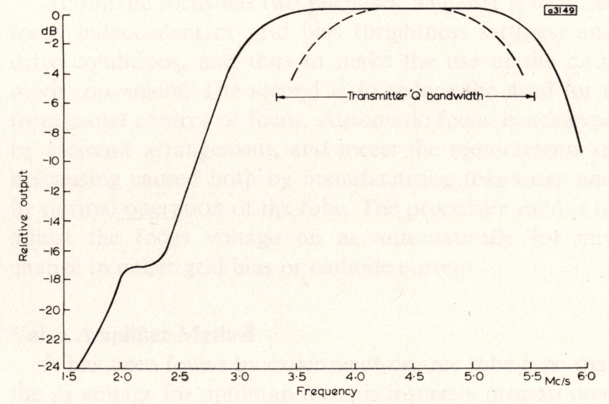


Fig. 17—Chrominance amplifier response curves
——— response to demodulator inputs
– – – – overall response to c.r.t. grid

SETTING UP

The alignment of the unit is quite simple; and, apart from a final adjustment of (B-Y) phase which is best carried out on a colour transmission, it can be readily accomplished with a c.w. signal generator covering the range 1·5 to 6Mc/s.

A suggested procedure is given below.

First remove line blanking, and ensure that d.c. operating conditions are correct. The signal is at all times connected to the input terminal.

Stage 1

Signal frequency	*Monitor*
1·57Mc/s	Top of L_2

Remove TR_1, and adjust L_1 for minimum output.

Stage 2

Signal frequency	*Monitor*
4·43Mc/s	Emitter TR_3

Replace TR_1, disconnect R_{11}, unscrew slug of L_7. Tune L_2, L_3, L_5 circuits for maximum output. Re-connect R_{11}. Depending on layout and circuit strays, it may be desirable to check that the response curve is perfectly flat, and to slightly retune L_3/L_5 as necessary.

Stage 3

Signal frequency	*Monitor*
6Mc/s	Emitter TR_3

Adjust L_4 for minimum output.

Stage 4

Connect reference signal, and align T_1 and T_2 as for normal bandpass circuit.

Set RV_2 in centre of range. Set c.r.t. drive controls to maximum.

Stage 5

Signal frequency	*Monitor*
100kc/s different from reference frequency	(B-Y) output

Adjust L_8 for clean 100kc/s signal. Set signal generator output to give voltage at (B-Y) output of 50V peak-to-peak.

Stage 6

Signal frequency and amplitude	*Monitor*
As in Stage 5	(R-Y) output

Adjust L_7 for clean 100kc/s output. If necessary, set RV_2 to give an (R-Y) output amplitude equal to 0·56 (B-Y). If T_1 and T_2 have been adjusted correctly, there will be a relative phase difference of 90° between the 100kc/s signal observed at the (R-Y) output and that present at the (B-Y) output.

Notes

The difference frequency of 100kc/s has been chosen quite arbitrarily, and any convenient frequency difference within the passband may be used.

A wobbulator can, of course, be used if preferred, in which case it is probably easiest to adjust the low-pass filters L_7 and L_8 on signal.

The only remaining adjustment is concerned with the picture tube drive ratios. Once the grey-scale tracking has been set up, the particular drive ratios for the tube in use will be known from the positions of the luminance drive controls. It is then necessary to set the colour difference drive controls in the same ratios. In the case of a sulphide phosphor tube such as the AX53-14, the colour difference drive ratios should be set so that the (R-Y) drive is approximately 10% high, in order to compensate for the difference between picture tube phosphor colours and those applying at the transmitter.

REFERENCE

Birt, D. R. and Freeman, K. G. 'An Appraisal of some Decoding Systems for Three-gun N.T.S.C. Colour Television Receivers'. **The Radio and Electronic Engineer,** Vol. 28, No. 1, July 1964, pp. 33 to 54.

Automatic Focus Circuit

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This article discusses means of maintaining focus, in electrostatic cathode ray tubes, by simple preset controls, for all drive conditions. Circuits for specific tubes are described.

INTRODUCTION

With many systems of displaying information, it is unnecessary to have controls relating to the display medium. For example, digital indicators or pen recorders fall into this category whereas the c.r.t. display requires controls for brightness, focus and astigmatism.

Controls are required both because of imperfections in the fundamentals of c.r.t. design and the unavoidable spread in manufacturing tolerances. It is desirable that the adjustments to controls for correcting these variations should be independent. Unfortunately, changes in the brightness setting cause changes in the required focus voltage in many modern tubes. The present article discusses ways* of avoiding this.

Astigmatism Control

The astigmatism control is often brought out as a front panel control on present day oscilloscopes. Its function is to compensate for defocusing fields set up between the deflecting systems and the beam-forming electrodes. These fields produce elongation of the spot in the x or y direction and are due to differences of potential existing between the beam-forming system and the mean values of the x or y deflecting systems. In any given situation, these potentials are normally fixed relative to each other and are not influenced by the operation of the tube. Theoretically, the only adjustment required is an initial one when the tube is first brought into service in the oscilloscope. However, it is usual to adjust astigmatism by varying the a_3 potential. This electrode carries a proportion of the cathode current. Unless a_3 is connected to a low impedance source, the voltage of which must be adjustable for astigmatism correction, variations in cathode current will produce voltage variations on this electrode. Thus only if a potentiometer of sufficiently low impedance is used for this purpose can it be preset.

Focus Control

The control of focus is necessary both because of manufacturing tolerance and because of effects caused in the tube when operating. Thus a preset control is not normally acceptable as many tubes which are designed to give a good focus (that is, small spot size) also have a focus voltage characteristic dependent upon the amount of beam current being drawn.

A third factor influencing the choice of an adjustable focus control is introduced in most tube designs by the presence of deflection-defocusing. This is usually an effect most noticeable in the x direction, whereby the edges of the display are out of focus. Adjustment of the focus enables those regions to be focused at the expense of centre focus, or permits some intermediate compromise setting.

Effect of Drive upon Focus

The variation of focus voltage required by changes of beam current are made less noticeable when the bright-up pulse is considerably smaller than the grid base cut-off voltage. For example, if the tube cut-off is −70V and the bright-up pulse is 30V, the normal range of grid bias will carry the pulsed grid voltage from −70V to −40V without the undeflected spot becoming visible. Thus only 43% of the available grid base is being used, and the defocusing is reduced at the expense of brightness.

To obtain the optimum performance from the c.r.t. it is necessary to provide a bright-up pulse of amplitude nearly equal to the cut-off of a tube on minimum limit. Thus, if the cut-off range is −45V to −90V, then the bright-up pulse should be between 40V and 44V. It is assumed that this criterion of design is met when giving consideration to the design of an automatic focus circuit.

TECHNIQUES TO OBTAIN AUTOMATIC FOCUS

Automatic focus has two purposes. The first is to make focus independent of grid bias (brightness settings) and drive conditions, and thus to make the use of the c.r.t. more convenient. The second is to reduce the need for a front-panel control of focus. Automatic focus is achieved by a circuit arrangement, and meets the requirements of defocusing caused both by manufacturing tolerances and by normal operation of the tube. The procedure used is to adjust the focus voltage on a_2 automatically for any change in either grid bias or cathode current.

Valve Amplifier Method

It has been found by experiment on one tube type that the a_2 voltage for optimum focus is inversely proportional to the cathode current. This relationship may be used by employing a large cathode resistance, making the tube act as a cathode follower. Such a method is undesirable because it causes changing sensitivity and increased drive requirements. Use may be made, however, of the relationship between cathode current and focus voltage by sampling the cathode current with a low value resistor and amplifying this voltage with a cathode-coupled double triode connected to a_2.

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*The material covered in this article is the subject of a patent application.

A sketch of the arrangement is shown in Fig. 1. A cathode-coupled double triode is used to amplify the voltage developed across R_3, and apply it to a_2. This two-valve arrangement is desirable with the d.c. coupling because it reduces the feedback which would be present with a single cathode-biased valve. The neon stabiliser provides a suitable voltage to enable a large value of R_6 to be used. R_4 adjusts the gain applied to a_2, and R_9 adjusts the d.c. level. Thus R_4 is adjusted at high brilliance, the condition of maximum drive to the tube, while R_9 is adjusted at low brilliance when little drive is applied. To be economical, the valve should run from the e.h.t. supply and draw a maximum of approximately 2mA. Under these conditions the bandwidth would be severely restricted and the circuit would be unsuitable for grid-modulated arrangements. With a beam-blanked tube the cathode current is substantially constant either blanked or unblanked, provided that grid or cathode modulation is not applied, so that this system may then be acceptable.

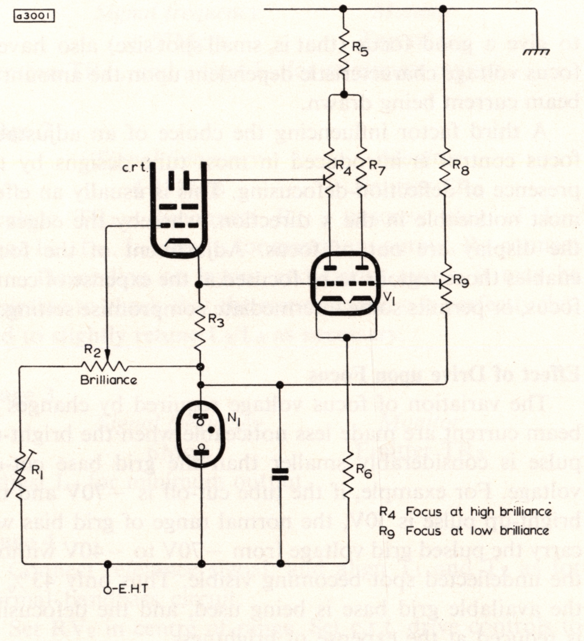

Fig. 1—Basic form of valve-amplifier automatic focus circuit

Ganged Potentiometer Method

The preferred method of continuously focusing the tube without the need of a valve involves the use of a double ganged potentiometer. With this system, one section of the potentiometer controls the brightness while the other corrects for the focusing action at the a_2 electrode. This technique has no frequency problem and can be used with either beam-blanked or conventionally operated tubes.

TUBE PROPERTIES ASSOCIATED WITH AUTOMATIC FOCUS

For the circuit to be designed correctly it is necessary to establish the various relationships between the control electrode voltage of the tube in relation to the condition of optimum focus. These properties may vary with the design of the tube gun; and certain tubes, such as split-beam types, have special problems associated with the effects of focus.

A fundamental property of any cathode ray tube gun is that as the point of focus is altered in space, the cone of electrons has a different angle of convergence from the focusing lens to the point of focus (Ref. 1). With the split-beam tube, the main cone of electrons is split into two smaller cones originating off the gun axis. Consequently, as the angle of convergence is altered by focusing, a shift in the y direction will take place on the screen. With the presence of defocusing caused by beam current changes, shift also takes place. Both these effects are undesirable.

APPLICATION OF AUTOMATIC FOCUS TO E10-11GH

To take full advantage of the automatic focus technique, E10-11GH split-beam tubes were selected for initial experiments. The curve of grid voltage versus a_2 voltage for optimum focus was plotted for three particular E10-11GH tubes chosen at random (Fig. 2). It is necessary to reproduce this relation by suitable adjustment to the law of the potentiometer controlling either of the electrodes.

A potentiometer network has a relation

$$V = nE \qquad \ldots(1)$$

where E is the voltage impressed across its ends, V is the output voltage and n is a function of the rotation, corresponding to the law of the potentiometer network. If we consider the circuit of Fig. 3, then the rotation of the potentiometer may be expressed as the fraction $r/(r+r')$. Now $r+r' = R$, the total resistance of the potentiometer, and the fractional rotation n_1 is r/R. That is, for a linear potentiometer, the rotation is a linear function of r, with R constant.

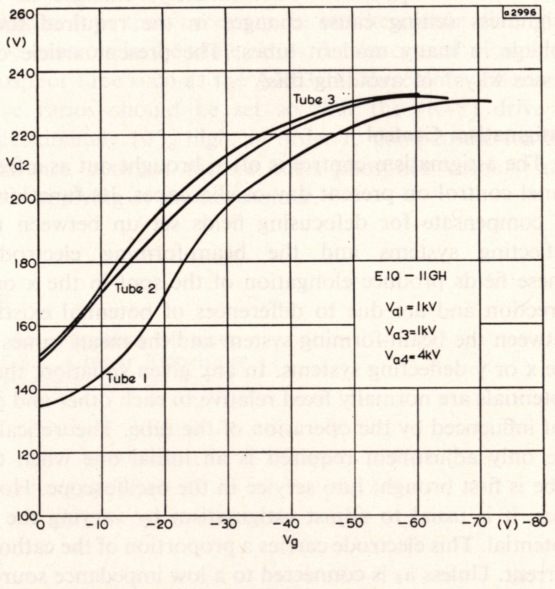

Fig. 2—Examples of curves of V_g versus V_{a2} for optimum focus, with the tube type E10-11GH. Three tubes were chosen at random

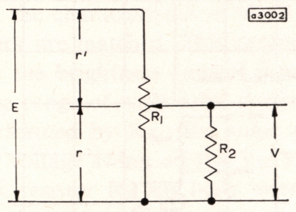

Fig. 3—First non-linear adaptation of linear potentiometer

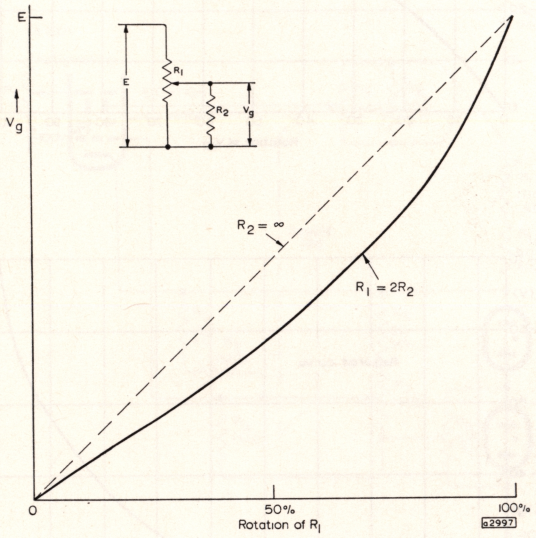

Fig. 4—Graph of output versus rotation for first adapted circuit

For the circuit of Fig. 3 the fractional rotation n_2 can be written

$$n_2 = \frac{\dfrac{r.R_2}{r+R_2}}{r' + \dfrac{r.R_2}{r+R_2}}. \qquad \ldots(2)$$

This equation can then be simplified to give

$$n_2 = \frac{r.R_2}{-r^2+r.R_1+R_1.R_2}. \qquad \ldots(3)$$

Fig. 4 shows the curve obtained from this equation for a particular case of $R_1 = 2R_2$. As R_2 increases the curve approaches the linear form.

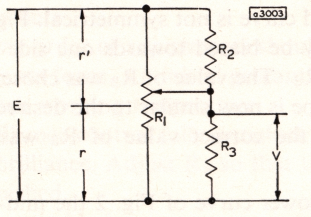

Fig. 5—Second non-linear adaptation of linear potentiometer

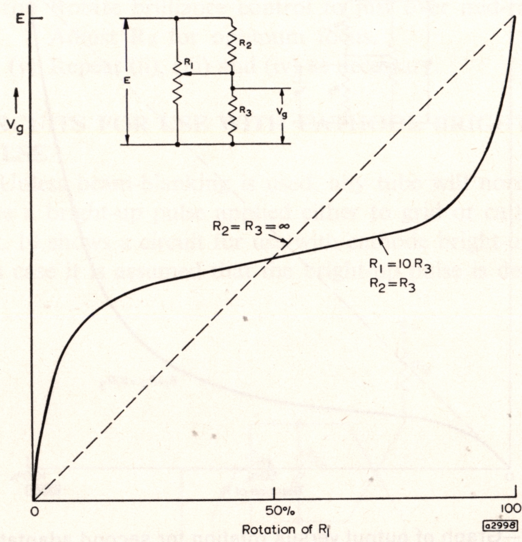

Fig. 6—Graph of output versus rotation for second adapted circuit with $R_2 = R_3$

To produce an S shape similar to the desired curves in Fig. 2 a further modification to the network is required. This may be obtained by adding resistance to the other arm of the potentiometer as in Fig. 5.

Then

$$n_3 = \frac{\dfrac{r.R_3}{r+R_3}}{\dfrac{r.R_3}{r+R_3} + \dfrac{r'.R_2}{r'+R_2}}. \qquad \ldots(4)$$

Simplification, as before, produces a value of n_3, the fractional rotation,

$$n_3 = \frac{-r^2.R_3+r(R_1.R_3+R_2.R_3)}{-r^2(R_2+R_3)+r(R_1.R_3+R_1.R_2)+R_1.R_2.R_3}. \qquad \ldots(5)$$

Fig. 6 shows the curve obtained with this circuit. Here the curve is symmetrical about the centre, since $R_2 = R_3$. As both R_2 and R_3 are increased in value, the curve approaches the linear form.

The desired curve is not symmetrical. Fig. 7 shows how the shape may be biased towards one side by making R_3 smaller than R_2. The value of R_2 was chosen to equal R_1. The basic shape is now similar to the desired curve (Fig. 2, Tube 1) and the correct value of R_3 was calculated as follows.

From the lower curve of Fig. 2 the mid-rotation point was found to be at 35% of the tube cut-off voltage. This represents 35% of the V_g scale on Fig. 7. Therefore, with

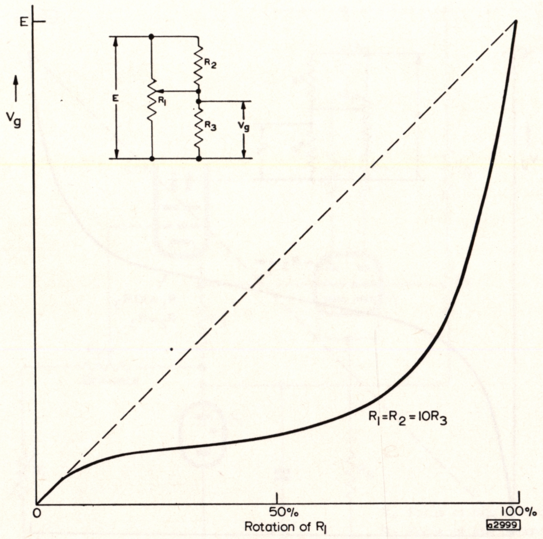

Fig. 7—Graph of output versus rotation for second adaptation with $R_2 \neq R_3$

the potentiometer at its centre setting (that is, 50% rotation), V_g must be 0·35E.

Then

$$\frac{\dfrac{0{\cdot}5 . R_1 . R_3}{0{\cdot}5 . R_1 + R_3}}{\dfrac{0{\cdot}5 . R_1 . R_3}{0{\cdot}5 . R_1 + R_3} + \dfrac{0{\cdot}5 . R_1 . R_2}{0{\cdot}5 . R_1 + R_2}} = 0{\cdot}35. \qquad \ldots(6)$$

Now, if we choose $R_1 = R_2$, as in Fig. 7, and for convenience let R_1 and R_2 be 1MΩ, then Eq. (6) provides a value of $R_3 = 0{\cdot}28$MΩ. Fig. 8 shows the experimental curve obtained using these calculated values. For comparison, the lower curve of Fig. 2 is redrawn and scaled to the same size. It may be seen from these two curves that a very close agreement has been obtained. A method of evaluating R_1 and R_2 without assuming a value, would be to select two points on the required curve and solve for R_1 and R_2 simultaneously.

AUTOMATIC FOCUS CIRCUIT

The circuit in Fig. 9 shows the final circuit as applied to the E10-11GH. The a_2 and grid potentiometers are shown correctly ganged so that full clockwise rotation of the potentiometer carries the sliders to the top of the resistors, as shown.

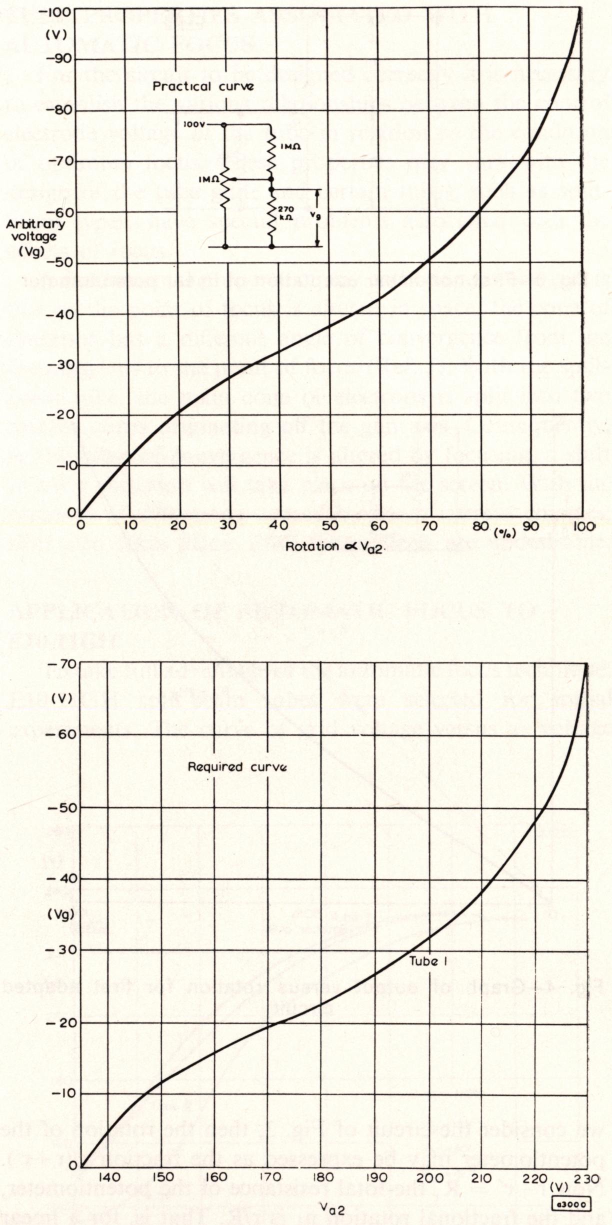

Fig. 8—Comparison of required curve of V_g versus V_{a2} and curve obtained in practical case

Use of Logarithmic Potentiometer

The curved characteristic of Fig. 8 shows that for a small rotation of the spindle a sudden change in V_g takes place near the cut-off region. This is undesirable as it makes the low brilliance settings rather 'fierce'. To overcome this, both gangs used are logarithmic. Thus the relation, one to the other, is linear before adding the shaping network, but the spindle rotation is logarithmic. This arrangement produces a satisfactory control over the complete range.

Function of Preset Controls

Resistor R_9 equalises the voltage across R_7 to that of the grid cut-off voltage. This is necessary as the voltage across R_7 (E of Eq. (1)) must be the same as the cut-off voltage so that the characteristics of the tube and potentiometer network are matched. This arrangement has the advantage that the brightness control functions over the whole grid base range of each tube. A similar equalising function is performed by R_8; the value of this resistor determines the voltage swing on the V_{a2} axis.

The variable resistor R_3 has been inserted to provide adjustment to the shape of the characteristic curve of Fig. 8. This allows for second order variations of curve-shape between tubes. In this way the three controls enable the amplitudes of each axis of Fig. 8 to be varied, as well as the shape of the curve. In addition, R_{11} enables the level of the a_2 voltage to be adjusted.

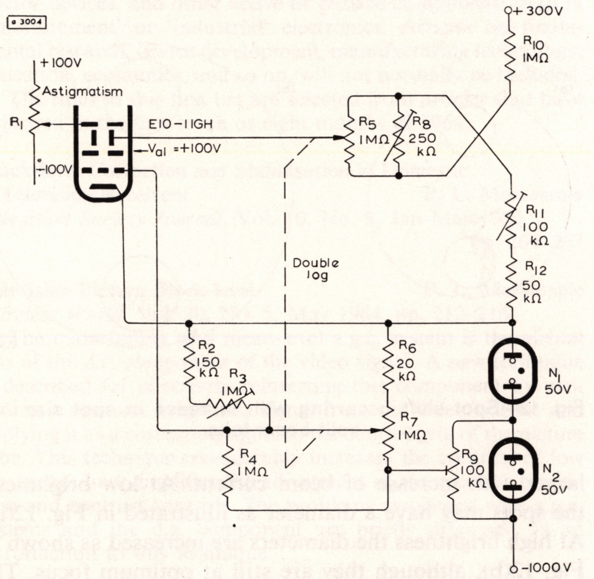

Fig 9—Final circuit applied to E10-11GH tube

Stabilisers

The neon voltage stabilisers N_1 and N_2 ensure that the voltage across $R_6 + R_7$ is constant for variations in the total resistance value of the network consisting of R_2, R_3, R_4 and R_7, such as occurs when R_7 is varied. They also stabilise against cathode current variation. Two 50V neon indicators are used to meet the requirements of low current, small size and low cost.

Passive Elements

The resistor R_{10} is the main voltage dropping resistor of the bleed chain and primarily determines the current flowing through the chain. In the present case the values chosen give a bleed current of 1mA with all preset resistors at mid-range. The resistor R_{12} is a 'hold-off' to allow a correct voltage range for R_{11}.

The introduction of R_2 into the correcting network gives smooth control with R_3 and prevents grid-to-cathode short circuit when R_3 is at zero before setting up; R_6 prevents the grid becoming more positive than −1V.

Setting-up Procedure

The following order of adjustments should be followed to obtain optimum focus. An initial precaution is to allow the tube to reach working temperature.

(i) Set brilliance control fully anticlockwise for minimum brilliance. Adjust R_9 so that the spot is just extinguished.

(ii) With a suitable low-occupance pattern displayed, rotate brilliance control fully clockwise.
Adjust R_1 (astigmatism control) and R_{11} for optimum focus.

(iii) Rotate brilliance control to give a dim trace.
Adjust R_1 and R_8 for optimum focus.

(iv) Rotate brilliance control to just over mid-range.
Adjust R_3 for optimum focus.

(v) Repeat (ii), (iii) and (iv) as necessary.

CIRCUITS FOR USE WITH CATHODE BRIGHT-UP PULSE

Unless beam-blanking is used, any tube will normally have a bright-up pulse applied either to grid or cathode. Fig. 10 shows a circuit for use with cathode bright-up. In this case it is assumed that the bright-up pulse is derived from a bistable circuit operating at cathode potential. In the circuit arrangement shown, the brightness potentiometer is connected at the maximum brilliance end to a potential of 140V–50V, or 90V, above the e.h.t. negative rail. The anode of the bistable valve V_1 stands at 140V during the tube 'off period', and pulses down to 95V during

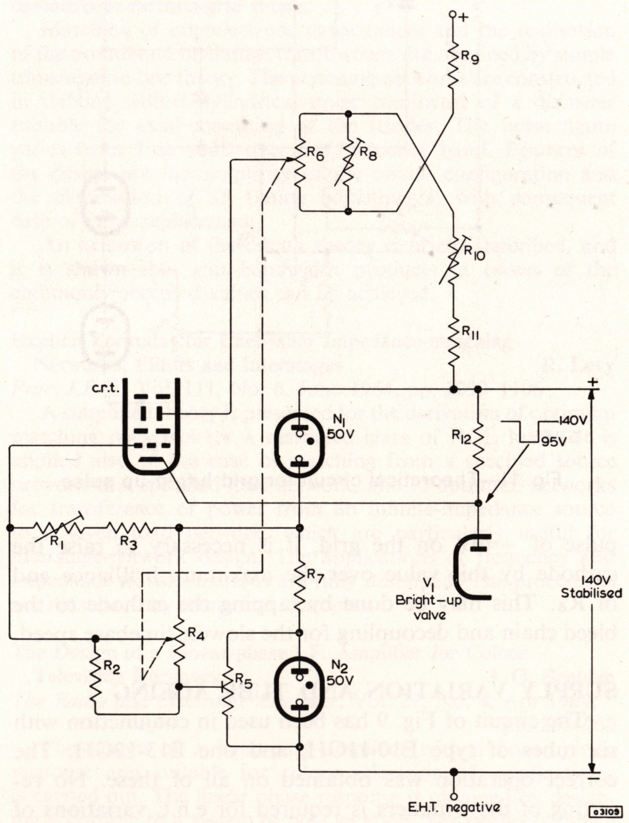

Fig. 10—Theoretical circuit for cathode bright-up pulse

the 'on period'. Therefore at maximum brilliance the grid is taken to -5V with the bright-up pulse applied. The cut-off is equalised, as before, with R_5.

GRID BRIGHT-UP

An alternative circuit using a grid bright-up is shown in Fig. 11. Here part of the bright-up pulse, divided by R_1 and R_2, is applied to the grid from the timebase circuit. The lower end of R_2 is returned to R_6 and this controls the grid voltage level. Assuming a bright-up pulse of $+40$V on the grid, it is necessary to raise the cathode by this value over the maximum brilliance end of R_6. This may be done by tapping the cathode to the bleed chain and decoupling for the slowest timebase speed.

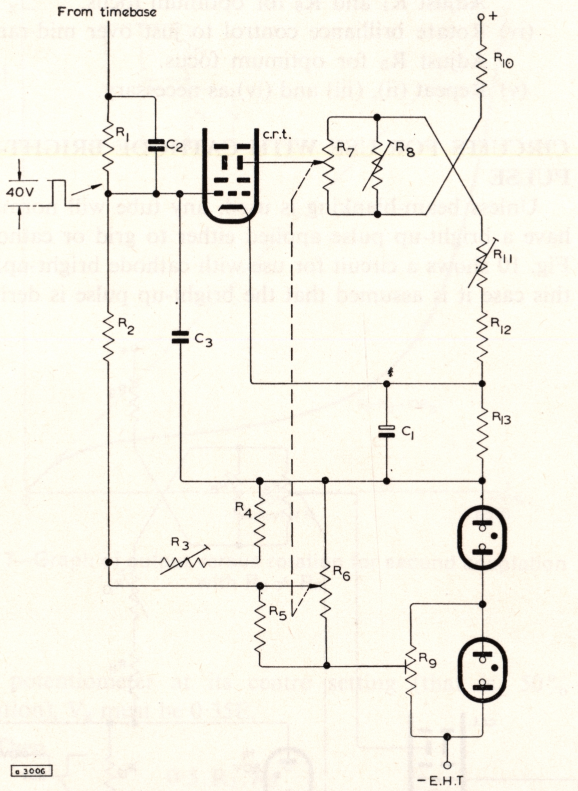

Fig. 11—Theoretical circuit for grid bright-up pulse

SUPPLY VARIATION AND TUBE AGEING

The circuit of Fig. 9 has been used in conjunction with six tubes of type E10-11GH, and one E13-12GH. The correct operation was obtained on all of these. No resetting of the trimmers is required for e.h.t. variations of up to $\pm 10\%$.

One E10-11GH was subjected to an artificial life ageing by reducing its heater voltage. The cut-off changed from -71V to -67V and the change in V_{a2} for maximum to minimum brightness (δV_{a2}) fell from $\delta V_{a2} = 108$ to $\delta V_{a2} = 54$V. This necessitated resetting all the preset controls, but once this was done automatic focus was again achieved.

PERFORMANCE

The circuit was used with the E10-11GH and the E13-12GH which are split-beam tubes. Some y shift is still present at near maximum brilliance setting although optimum focus is maintained. This may be attributed to the fact that the spot has a finite size, the size becoming larger with increase of beam current. At low brightness. the spots may have a diameter as illustrated in Fig. 12(a), At high brightness the diameters are increased as shown in Fig. 12(b), although they are still at optimum focus. The outer circle illustrates the area of the spot in the absence of a split beam plate. Then for the larger spot area, it can be seen that the centre of the split spots are a greater distance apart than for the smaller area spot. Hence some line movement has occurred. This effect is slight compared with the shift encountered without the automatic focus circuit. The E13-12GH shows much less defocusing with drive than the E10-11GH. Hence the need for automatic focus is less. The circuit was also used with the D13-22GH, a single-beam mesh tube, and found to be satisfactory.

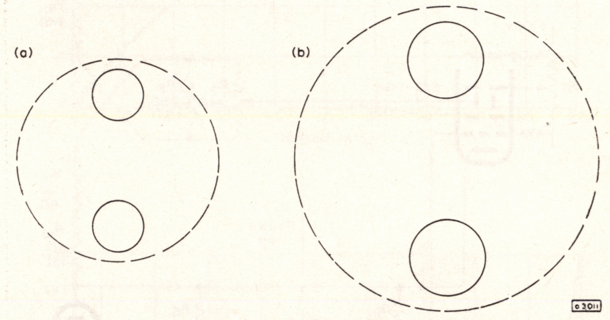

Fig. 12—Spot-shift occurring with increase in spot size in a split-beam tube

In all cases the circuit involves the use of additional components. These can, however, include inexpensive preset potentiometers and simple neon indicator tubes. A ganged potentiometer is required but a front-panel focus control can be saved.

An additional advantage of the circuit is that the full brightness control range can be used on every tube.

REFERENCE

1. GRODSZINSKY, E. J. 'Dual Trace Oscilloscope Tubes', **Mullard Technical Communications**, Vol. 7, No. 66, August 1963, pp. 179 to 183.

Articles by Mullard Authors

From time to time we shall publish a list of articles of Mullard origin that have appeared in other journals. The articles will, in general, be of the same nature as those published in *Mullard Technical Communications;* that is, they will be concerned with applications of valves, cathode ray tubes, semiconductor devices, and other active or passive components used in 'entertainment' or 'industrial' electronics. Articles on fundamental research, device development, manufacturing technology, education, economics, and so on, will not normally be included.

The titles in this first list are selected from articles that have appeared in the first seven or eight months of 1964.

Black-level Correction and Stabilisation in Domestic Television Receivers **P. L. Mothersole**
Television Society Journal, Vol. 10, No. 9, Jan-Mar 1964, pp. 264–267

Television Picture Black-level **P. L. Mothersole**
Wireless World, Vol. 70, No. 5, May 1964, pp. 212–216

The main failing of a mean-level a.g.c. system is the virtual loss of the d.c. component of the video signal. A new technique is described for effectively reinserting this component by producing a separate brightness signal from the video output and applying it as a correction signal to the control grid of the picture tube. This technique considerably increases the contrast of low key scenes and simplifies the adjustment of the receiver brightness and contrast controls. The problems associated with a.g.c. systems and the d.c. component are briefly discussed as an introduction to this technique.

Television Receiver Design Trends - A Dual-standard Hybrid Receiver **P. L. Mothersole**
Television Society Journal, Vol. 10, No. 9, Jan-Mar 1964, pp. 268–277

The drive requirement of the picture tube in a domestic television receiver limits the application of transistors to the small signal circuits and makes the fully transistorised receiver only practicable at the moment with small cathode-ray tubes operating at low e.h.t. potentials. The tuner, i.f. and audio amplifiers can be transistorised, and using the new range of transistor types now available a performance similar or superior to that currently obtained with valves can be achieved. A hybrid receiver employing transistors for the small signal circuits and valves for the power circuits is a practical proposition. Some of the special problems encountered in such a receiver are described and possible solutions discussed.

Protecting Aircraft Distribution Systems **K. F. Bacon**
Industrial Electronics, Vol. 2, No. 7, July 1964, pp. 334–338

Busbar protection is now highly important in aircraft, but conventional methods are expensive and heavy with the three-phase supplies now widely used. This article describes a simple system which depends on the detection of the positive, negative and zero sequence components of voltage which appear when the supply is unbalanced.

Characteristics of Broadband Parametric Amplifiers using Filter Networks **B. L. Humphreys**
Proc. I.E.E., Vol. 111, No. 2, Feb 1964, pp. 264–274

The paper gives a derivation of the relation between the bandwidth and the gain of a single diode parametric amplifier using filter networks at both signal and idling frequencies. Approximate algebraic expressions are obtained and compared with computed characteristics. The effects of small variations in component values are computed in an attempt to estimate the probable permissible tolerances for diode parameters and circuit elements. The phase- and group-delay characteristics are also computed, and the latter is shown to be of the order of twice the inverse of 2π times the amplifier bandwidth.

The noise contribution due to the parametric action is shown to be a function of the signal frequency gain and the signal/idling frequency ratio. Because this relationship is such that the noise contribution falls with decreasing gain, the effective bandwidth of a low-noise receiver using a parametric amplifier as the first stage may be substantially greater than the normal bandwidth defined by a reduction of 3dB in the gain.

Reported practical results are briefly examined in the light of the theory.

Wideband U.H.F. Amplifier using Earthed-grid Triodes **R. Levy, S. J. Robinson, and S. K. Salmon**
Proc. I.E.E., Vol. 111, No. 2, Feb 1964, pp. 235–240

An untuned u.h.f. amplifier giving a mean gain of 25dB from 650 to 1000Mc/s is described. Three disc-sealed triodes are cascaded in earthed-grid stages.

Matching of interelectrode capacitances and the realisation of the broadband interstage transformers are obtained by simple transmission-line theory. The matching networks are constructed in slabline with a cylindrical outer conductor of a diameter suitable for axial mounting of the triodes. The noise figure varies from 7 to 10dB over the frequency band. Features of the design are the simplicity of the circuit configuration and the elimination of all tuning adjustments, with consequent ease of valve replacement.

An extension of the circuit theory is briefly described, and it is shown that gain-bandwidth products in excess of the commonly accepted values can be achieved.

Explicit Formulas for Chebyshev Impedance-matching Networks, Filters and Interstages **R. Levy**
Proc. I.E.E., Vol. 111, No. 6, June 1964, pp. 1099–1106

A simplified theory is presented for the derivation of optimum matching networks for a restricted class of RCL loads. It is applied also to the case of matching from a specified source network to a specified load network, and to optimum networks for transference of power from an infinite-impedance source to a given load network, which are particularly useful for interstage-network design. The applications of frequency and impedance transformations are fully described. The procedures are illustrated by numerical examples.

The Design of a Linear-phase I.F. Amplifier for Colour Television Receivers **J. O. Scanlan**
The Radio and Electronic Engineer, Vol. 27, No. 4, Apr 1964, pp. 255–260

The design of a band-pass amplifier with a linear phase response and suitable for colour television i.f. applications, is carried out. The linear phase response is obtained by designing according to a system of pole location. The results obtained from tests on the i.f. amplifier are given, and are found to be in good agreement with the theory.

Cross-colour and Luminance Twinkle in N.T.S.C. Colour Receivers with Various Display Devices **K. G. Freeman**
The Radio and Electronic Engineer, Vol. 27, No. 4, April 1964, pp. 247–254

It has been observed that under some conditions N.T.S.C. colour receivers employing single-gun tubes exhibit severe cross-colour and luminance twinkle due to the presence of high video-frequency luminance information and noise. From analysis of a simple case it is established that this effect is the result of sampling of the high-frequency luminance and therefore noise components by the colour selection process (necessary to decode the subcarrier chrominance information) which takes place at the screen. Methods of eliminating or substantially reducing this objectionable effect are discussed.

Electron Beam Machining **J. Kelley**

Brit. Communications and Electronics, Vol. 11, No. 1, Jan 1964, pp. 20–23

A high energy beam of electrons can be focused into a small spot, providing a source of extremely high energy density. With this source as a tool any material can be cut or drilled. The technique is described and the present state of technology discussed. The theoretical limitations are tested and some applications outlined.

An Image Display for a Mechanical Photo-multiplier Scanner **H. A. Dell**

Brit. Communications and Electronics, Vol. 11, No. 3, Mar 1964, pp. 172–175

A simple mechanical flying spot scanner suitable for testing the uniformity of the cathodes of photo-multipliers can be made with only two rotating lenses. With a tungsten filament light source this instrument is photometrically both controlled and predictable. A cathode-ray tube display for the scanner has now been developed, which produces an image representing in light and shade the features of the photo-multiplier cathode explored.

Correlation Measuring Equipment for a Stellar Interferometer **A. Browne**

Brit. Communications and Electronics, Vol. 11, No. 7, July 1964, pp. 472–475

The measurement of the angular diameter of optical stars is a major problem in astronomy. The largest telescopes can just resolve the largest stars, but atmospheric fluctuations made accurate measurement impossible. This also applies to the Michelson optical interferometer which effectively increases the resolving power of the telescope. In the case of radio stars, however, measurements of the angular diameter have been made by means of the intensity interferometer, and it was proposed that a similar interferometer could be built for optical stars. This led to the construction of the intensity interferometer for optical stars which is described in this article.

An 85Mc/s Transistorised Transmitter **A. E. Hilling**

Electronic Engineering, Vol. 36, No. 437, July 1964, pp. 466–469

An 85Mc/s transmitter using the AFY19 alloy-diffused germanium transistor is described. This transmitter produces a peak envelope power of 1W when 100 per cent amplitude modulated.

This circuit would be suitable for low power portable transmitters in the 71·5 to 88Mc/s band.

A Logarithmic Time-base **R. N. Alcock**

Electronic Engineering, Vol. 36, No. 436, June 1964, pp. 404–406

The design of a passive logarithmic function generating network is described. A step voltage is applied to the input and the output increases with the logarithm of time. The network consists of a number of integrators with differing time-constants. The time range for logarithmic operation increases with the number of integrators. An experimental one-shot time-base was constructed which operated between 22msec and 30sec to an accuracy of 2 per cent.

A Simple Electronic Milliameter **G. F. Weston and P. Schagen**

Electronic Engineering, Vol. 36, No. 431, Jan 1964, pp. 21–23

An experimental glow discharge indicator is described in which the length of the glowing column is a linear function of the current. The length can be read with an accuracy of 5 per cent, and thus the tube can be used in applications where an indicating current meter is at present employed. The parameters affecting the full scale range are given. The tube is processed with the aid of standard sputtering techniques ensuring good initial and life characteristics.

Transistors in Television **D. G. Thompson**

(1) Portability, Reliability and Stability
Electrical and Radio Trading, Vol. 62, No. 791, 7 May 1964, pp. 59–60

(2) Valve and Transistor Circuits Compared
ibid, No. 792, 14 May 1964, pp. 35–39

(3) The Requirements are Exacting
ibid, No. 793, 21 May 1964, pp. 35–36

(4) I.F. Amplifiers
ibid, No. 794, 28 May 1964, pp. 39–40

(5) V.H.F. Tuners
ibid, No. 797, 18 June 1964, pp. 39–41
